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CONTENTS

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FOREWORD

THIS booklet is based on a series of six articles written by one of us and published in the *R.S.G.B. Bulletin* during 1943. At that date information on several branches of microwave technique could not be published owing to the secrecy regulations. The articles, therefore, have been largely expanded, rewritten and brought up-to-date and an equal amount of entirely new material added.

The approach to the subject is mainly theoretical, although in a few cases practical information is given, as it is still too early to indicate in what ways the amateur can experiment at microwave frequencies. When the commercial availability of specialised components is known, and some experience of amateur apparatus has been gained, it is hoped to publish a companion booklet which will deal entirely with the practical side. This is not to say that the enterprising experimenter cannot use microwaves at the present; by the exercise of ingenuity and following the general principles laid down in this present publication, a lot can be done. It is hoped that those readers fortunate enough to be able to carry on microwave work will continue and collect material which can be incorporated in the companion booklet. A publication of this kind is always a co-operative effort and it is only by summarising and correlating the experiences of others that a comprehensive review can be presented.

We wish to express our indebtedness to all those who have helped in the preparation of the booklet and particularly to the following firms who have given permission for the use of their photographs as illustrations: *The British Thomson-Houston Co. Ltd., Electrical and Musical Industries Ltd.,* and *Marconi Instruments Ltd.*

J. H. S.
E. D. H.

CHAPTER 1. INTRODUCTION TO MICROWAVES

Frequencies

THE band of frequency which may be termed microwaves is variously defined, but a reasonable definition is that it covers the range 500–25,000 Mc/s. (60 cms.–1·2 cms.). The lower limit is chosen because it is about the upper limit of operation of ordinarily constructed valves with lumped element circuits, and 25,000 Mc/s. because it is the highest band at present usefully employed. Frequencies above that figure have been generated and used in some applications but this review will not consider them. Between 500–1,000 Mc/s. very little work is done as it is the Cinderella of the microwaves, being too high for ordinary techniques and a little too low for special microwave methods. Above 1,000 Mc/s. the klystron, and particularly the resonant cavity magnetron, come into their own; the wave guide becomes of practicable size, and the receiver becomes a relatively simple affair using a crystal mixer.

Until comparatively recently the frequency spectrum above 500 Mc/s. has been unoccupied except for a few experimenters. Commercial installations operating in this part of the spectrum were practically non-existent, the cross-Channel microwave link on 17 cms. being an exception. This was due to the fact that there was no method of generating adequate power at these very high frequencies. At the time of installation of the cross-Channel link the most satisfactory method of producing oscillations at frequencies exceeding, say, 600 Mc/s. was by means of the positive-grid triode, which at best is only capable of generating a few watts of power, and with a very low overall efficiency.

However, with the development of new types of vacuum tube devices it is now possible to generate several hundred watts at frequencies of the order of 3,000 Mc/s. with greatly increased efficiencies, and a vast new frequency-spectrum has been opened for commercial activity.

As will be apparent from later chapters, a peculiarity of many microwave components is that, by virtue of their construction and mode of operation, their working frequency is fixed by their dimensions and once made cannot be altered except by some few per cent. Microwave components are, therefore, pretuned and have to be chosen with due regard to the band of frequencies which is desired to use, for instance, a 10 cms. klystron cannot be coupled to a 3 cms. wave guide! Such pretuned circuits and components are termed "preplumbed"—from the fact that the "plumbing" is fixed by the design. Two bands which have been extensively used are the 10 cms. (3,000 Mc/s., S-band) and the 3 cms. (10,000 Mc/s., X-band) bands, both for radar and communication. The 3 cms. band is standardised for the marine radar equipment which is now being made in large quantities and preplumbed equipment for this wavelength should soon be generally available. Although, speaking generally, the bands are named according to the wavelength, within any one band it is usual to specify frequency and quote band limits, etc., in megacycles per second.

Propagation

Normally the range of any microwave radiating system is only a little more than the optical range, although there is a certain amount of diffraction over the optical horizon. When considering very short wavelengths the field strength at a distant point is the resultant of at least two waves. There is

the space wave travelling directly from the transmitter to the receiving aerial, and there is also a ground-reflected wave reaching the receiver by a longer path (Fig. 1). If the aerials are very low there is also a wave which is more or less guided along the surface of the ground and if there are any obstructions, such as buildings, in the immediate vicinity there will be reflected waves from the walls of the obstructions. As all these paths will be of different lengths the phases of the different received waves will be different and so a series of " standing waves " will be set up, *i.e.* by moving the receiving aerial through several wavelengths, several maxima and minima of reception will be observed depending on whether the reflected waves arrive in or out of phase with the space wave. When a wave is reflected by a surface, there is a phase-change of 180°.

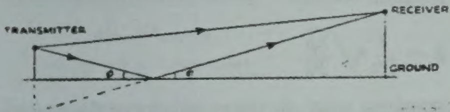


Fig. 1.
Direct and reflected wave paths between transmitter and receiver.

Now if we consider reception over flat ground with no lateral obstructions, the received signal is the resultant of the space and reflected waves, so that the reflected wave will arrive nearly 180° out-of-phase with the space wave and cause destructive interference. As the path lengths differ, however, the two will not entirely cancel out. Thus, the range over flat ground may be expected to be small if the aerials are low. If the aerials can be raised high enough to minimise the adverse effect of the reflected wave the range is very much increased.

Atmospheric conditions, such as rain and fog, have an effect on propagation and may cause bad fading if the range is great. For this reason microwave communication and radar sets cannot be reliably used in, say, tropical countries where sudden torrential downpours are experienced. Fog is not so serious in its effect and indeed radar sets find some of their greatest usefulness under foggy conditions.

CHAPTER 2. CIRCUIT ELEMENTS—TUNED CIRCUITS AND CAVITY RESONATORS

TO the uninitiated the R.F. portions of a microwave set appear confusing and incomprehensible. No coils or condensers are visible, perhaps even no valves; instead there is an arrangement of pipes, cables and boxes, with sliding elements for tuning and control. These components are, however, only higher-frequency counterparts of the more familiar circuit elements employed in lower frequency equipment, and their operation and construction is based upon normal electrical principles.

Resonant Circuits and Lines

Perhaps the most important part of a R.F. circuit is the resonant or tuned circuit and the reasons for the abandonment of ordinary types will now be briefly considered.

When a normal resonant circuit, consisting of an inductance in parallel with a capacitance, is adapted so as to tune to wavelengths in the microwave

spectrum, it is found that, as the wavelength approaches the dimensions of the circuit, there will be an increasing amount of power radiated by the circuit, and hence lost. This radiation loss is one of the major problems confronting the centimetre-wave engineer.

The usual way of dealing with a resonant circuit at the lower frequencies is to consider the coil as a pure inductance in series with a resistance, this latter being the ohmic resistance of the coil with an allowance made for the skin effect. No allowance is made for the loss due to radiation, which is negligible at the lower frequencies. As the frequency is increased the current in a conductor tends to flow more and more on the surface layer. For example, at a wavelength of 3 metres (100 Mc/s.) the thickness of the current-carrying layer in a copper conductor is less than 0·001 inch. The thickness of the layer may be found from the formula:—

$$\delta = \frac{1}{2\pi}\sqrt{\frac{\lambda\rho}{30}}$$

where λ is the wavelength in centimetres, and ρ is the resistivity of the material in ohms per cubic centimetre. Thus the thickness of the conducting layer is proportional to the square root of the wavelength. At centimetre wavelengths the conducting layer is extremely thin and the current should be regarded as flowing on the conductor instead of in it, and giving rise to electric and magnetic fields in the vicinity of the conductor. These fields travel away from the source with the velocity of light and any electro-magnetic field leakage from the circuit will constitute a radiation loss.

In an attempt to reduce this radiation loss, transmission lines were used as resonant circuits, the best known being the quarter-wave Lecher system consisting of two parallel wires or tubes a quarter wavelength long and short-circuited at the remote end. Centimetre waves are guided along these rods from the source to the remote end where reflection takes place at the short-circuit. The fields extend for a considerable distance into the medium around the conductors, and due to imperfections of the short-circuit, total reflection does not take place and a certain amount of electro-magnetic energy escapes and is lost from the circuit. The electro-magnetic energy should be regarded mainly as being in the dielectric medium in the vicinity of the conductors, and not penetrating the conductors to any great extent. The conductors are merely boundaries between which waves are propagated.

Since the electric and magnetic fields do not penetrate metal to any great extent, then resonant circuits for use on centimetre wavelengths should be self-enclosing so that the fields do not extend outside the circuit and thus the radiation loss from the circuit is zero. These conditions are fulfilled by a half-wavelength of concentric transmission line shorted at both ends. Consider a length of concentric line transmitting high-frequency energy from the point of view of electro-magnetic field distribution, instead of adopting the concept of current electricity usually applied to circuits operating on lower frequencies. Fig. 2 shows a cross section and a longitudinal section of a length of concentric transmission line. No field extends outside the line and hence radiation is zero. The lines of electric force are entirely radial and the regions of maximum and minimum electric flux-density are regularly spaced along the length of the line. The lines of magnetic force are concentric with the line and have regions of maximum density where the electric flux-lines have regions of minimum density. Neither the electric nor the magnetic field have components in the direction of propagation of the energy.

A self-enclosing resonant circuit may be made up of any even number of

half-waves, short-circuits being provided at points along the line (such as A and B in Fig. 2) where the electric field is zero. A resonant circuit, consisting of one or more half-wavelengths of line short-circuited at the ends, will be non-radiating since the fields will be entirely contained by the circuit. The waves should be thought of as travelling backwards and forwards along the length of line, reflection taking place at the short-circuits (voltage nodes).

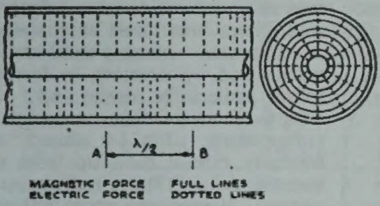


Fig. 2.
Field configuration in a coaxial line.

For maximum voltage step-up ratio the length of the line should be one half-wave long. Actually the step-up ratio is maximum for a quarter-wavelength line, open at one end, but such a quarter-wave circuit will radiate unless the open end is shielded (as in Figs. 18 and 19, page 23). Such half- and quarter-wave concentric circuits are suitable for use down to about 5 cms., provided that the inside diameter of the outer conductor is reduced with the wavelength. The reason for this will be apparent when cavity resonators have been discussed. Coupling to these circuits is usually effected by inserting a small loop at the short-circuited end where the magnetic field is at a maximum (or at any voltage node in the case of a length of line more than one half-wave long).

Butterfly Circuits

If it is desired to cover a wide frequency range at frequencies between, say, 300 and 1,000 Mc/s. then the resonant-line type of circuit becomes rather cumbersome and mechanical problems are encountered in tuning, and the ordinary type of tuned circuit does not give a high enough *Q*. For these frequencies, which, as stated in Chapter 1, are a little awkward to deal with, a compromise has to be adopted and a circuit used which combines the good points of both the high- and low-frequency types. A new type of circuit has been developed, known as the "butterfly," due to the shape of the condenser vanes. This consists of a split-stator condenser and two inductance straps in parallel, one on each side of the condenser assembly (Fig. 3). The tuning range of this circuit can be made very wide (of the order of three octaves), as when the condenser vanes are not meshed the inductance of the two straps is reduced due to the proximity of the rotor vanes, so that movement of the condenser vanes varies both the capacity and inductance of the circuit. A reduction of inductance of about one-third is claimed. The stator vanes and inductances may be supported at the electrical midpoints AA (Fig. 3) thus removing any solid dielectric from high-frequency fields. The valve electrodes should be connected to the tips of the stator vanes BB which are the high potential points in the circuit. As the fields are confined mainly to the inside of the circuit, radiation losses are not excessive. The number of vanes, however, should be kept to the minimum required to give the desired range, otherwise trouble may be encountered from unwanted longitudinal resonances. Output from the circuit may be taken from a loop coupled to the low-potential part of the inductance straps, or by tapping symmetrically

on to one of the inductances. A disadvantage, of course, is that the complete range is covered in only 90° rotation of the tuning spindle, although other assemblies are possible, using this same principle of inductance reduction, in which the range is spread out over 270°.

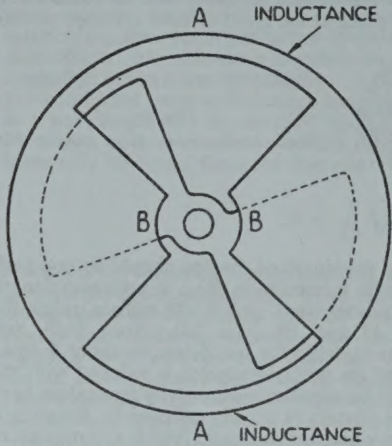


Fig. 3.
Tuning element of butterfly circuit.

The butterfly circuit permits the construction of a very compact wide-range oscillator suitable for use as a receiver local oscillator, and it has been used extensively in equipment—notably test instruments and wavemeters, in forms either especially made up or adapted from existing commercial components. Fig. 4 shows a butterfly circuit built up from a commercial split stator tuning condenser. It uses two acorn valves in push-pull as oscillators and broad strips of metal connected across the stator vanes as the inductive element. These points and the four filament chokes can be clearly seen in the illustration.

Cavity Resonators

Before proceeding further it should be pointed out that for a proper understanding of the operation of cavity resonators and wave guides a fairly high standard of mathematics is necessary, involving the use of Maxwell's field equations. Such a mathematical treatment is beyond the scope of this booklet. For those who are interested in the mathematics of the subject there are available numerous papers, and some text-books dealing exclusively with the theoretical treatment of the subject.

Cavity resonators should perhaps be considered with wave guides, which are covered in the next chapter, but for the sake of completeness are here classed under tuned circuits. The techniques associated with the two subjects are precisely similar and knowledge of one will help with the understanding of the other. The essential difference between them is, of course, that whilst

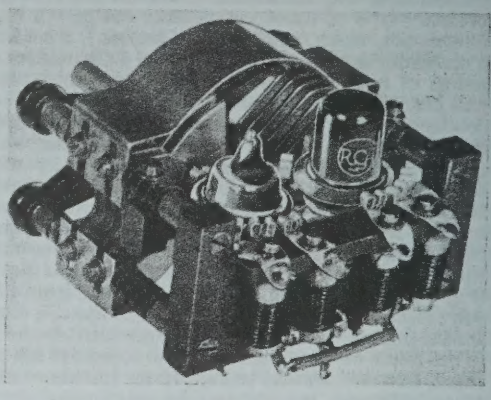


Fig. 4.
Practical butterfly circuit. (*Photo : courtesy of B. T.-H. Ltd., Rugby.*)

resonators are used in conjunction with other circuit elements for tuning purposes and thus serve a primary function in the operation of microwave equipment, wave guides are only used to pipe—literally to pipe—R.F. energy from one point to another and often only serve as items of mechanical convenience. They have, however, their own functions in impedance changing, matching, etc.

The concept of resonances in hollow bodies will appear strange to those accustomed to dealing with more conventional resonant circuits, but there is a way of deriving a cavity resonator from the more conventional type of resonant circuit.

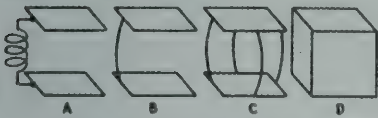
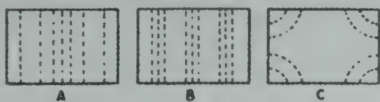


Fig. 5.
Derivation of cavity resonator from the conventional tuned circuit.

Consider a resonant circuit consisting of two condenser plates tuned to resonance by means of an inductance (Fig. 5A). In order to increase the resonant frequency of the circuit the size of the inductance must be reduced (keeping the condenser unchanged) until it consists merely of a heavy metal strip connecting the two plates (Fig. 5B). To increase the frequency still further, several such strips may be connected in parallel to give a lower inductance. These may be connected around the condenser plates as is shown in Fig. 5C. If the inductance is still further decreased by connecting an infinite number of conductors in parallel around the edges of the condenser

Fig. 6.
Modes of oscillation in a rectangular resonator.



plates we obtain a closed box (Fig. 5D). If such a hollow box is excited at its resonant frequency the top and bottom of the box will become charged and a high current will flow up and down the sides of the box.

The electric field between the top and bottom of the box will not be uniformly distributed but will be the most intense at the centre and will fall off at the edges to a very low value. Fig. 6A shows the electric field distribution for the simplest case of resonance in a rectangular box resonator. From the field distribution it will be seen that for the box to resonate at a given wavelength there must be a half-wavelength (or a multiple of half-wavelengths) across the face of the box. Thus cavity resonators will have a definite cut-off frequency (determined by the cross-sectional dimensions) below which the circuit will not resonate.

There are other modes of oscillation in the box resonator under consideration. Fig. 6B shows a case in which there are three half-waves across the face of the box and in Fig. 6C a mode is shown in which it is not the top and bottom of the box which resonate against each other but the sides which resonate against the top and bottom.

In Fig. 7 the field distributions in a hollow sphere A and in a toroidal rhumbatron resonator (" doughnut ") B are shown. This latter is important as it is much used in the klystron and other oscillators. The electric field is very intense across the neck through which the electron beam passes thus ensuring deep velocity modulation of the beam, and the construction of the resonator is such that the distance traversed by the electron beam is as short

as possible, thus securing a very small transit angle without resorting to excessively high accelerating potentials in the valve. Actually, any cavity resonator could be used in a velocity modulated oscillator provided that a sufficiently high accelerating potential could be obtained. In some cases a resonator with a transit angle of π radians, or multiples of π radians may be used but the toroidal rhumbatron has been found to be the most convenient.

Since the circuits described are self enclosing there will be no radiation loss, and as there is a very large metal surface over which the oscillating currents flow, the Q of a cavity resonator will be very high, being given approximately by the formula:—

$$Q = \frac{\text{Volume of inner space}}{\text{Surface area } x\ \delta}$$

where δ is the thickness of the inner conducting layer of the wall, and which may be calculated from the equation given previously. Values of Q of the order of 50,000 may be obtained.

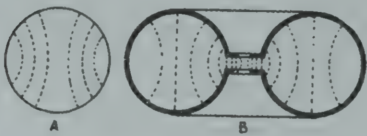


Fig. 7.
Field configuration in (A) a sphere, (B) a toroidal rhumbatron.

If the thickness of the metal enclosing the cavity is appreciable, *i.e.* thick compared to the depth of penetration of the current, points on the outside surface of the resonator will be at zero radio-frequency potential and if it is desired to measure the radio frequency voltage between two points (say between the top and bottom of a rectangular box cavity), then the measuring device must be connected to the inside surfaces of the resonator. If connection is made to the outside surfaces there will be no indication of any radio frequency voltage. Thus the inside and outside surfaces cannot be regarded as being of the same potential.

Coupling to cavity resonators is usually effected by means of a small loop inserted in the resonant space at a point where the magnetic field is most intense, but the resonator may also be excited by inserting a small probe along a line of electric force. The problem of coupling to hollow resonators will be treated in the next chapter.

Each cavity will have several modes of oscillation and thus may resonate at several frequencies so that care must be taken to introduce the electromagnetic energy at the correct point so that the desired oscillation mode may be set up. Cavities may be used as filters and acceptor circuits in the same way as conventional resonant circuits. Tuning is usually effected by making one of the walls in the form of a movable piston with finger contacts pressing on the other walls. As the Q of these circuits is usually very large, the tuning will be very sharp, necessitating a screw-thread device for moving the tuning piston.

In the case of the toroidal shaped resonator, the piston-tuning is impossible, but the frequency of the resonator may be altered, either by making the walls compressible or by introducing a plug of metal into the interior of the rhumbatron, thus altering the volume of the resonant space. These points are further illustrated with reference to the klystron valve in Chapter 4.

CHAPTER 3. CIRCUIT ELEMENTS—WAVE GUIDES

As long ago as 1897 Lord Rayleigh proved mathematically that electro-magnetic waves could be propagated along hollow tubes, without having a return conductor as is usual in the case of a radio frequency transmission line. This discovery was not, however, utilised until about 1935, due to the fact that the cross-sectional dimensions of the hollow tube had to be of the same order of dimensions as the wavelength of the waves to be transmitted, and as there were no means of generating centimetre waves the large size of the tube necessary for the wavelengths then in use made the system impracticable. However, in 1936 when the magnetron and Barkhausen oscillators were available for the production of energy at wavelengths down to 10 cms., interest in " wave guide " transmission lines revived and several theoretical and experimental papers on the subject were published in the U.S.A.

At centimetre wavelengths the distance between two components in a piece of apparatus may be a large number of wavelengths so the problem of transmitting energy from one point to another, without radiation loss, becomes very important. Here again, completely enclosed transmission lines must be used. The coaxial line is the commonest type of non-radiating transmission line and this may be used down to very short wavelengths provided that there is an appreciable thickness of metal in the outer conductor, and that the system is closed at each end.

There is one other requirement when using a coaxial line at very short wavelengths—that is, to make sure that the dimensions of the line do not become comparable with the wavelengths transmitted. The reason for this will become apparent when the hollow tube transmission line has been considered later in this chapter. If these conditions are observed then the coaxial line will behave as it does at lower radio frequencies and its operation in this manner need not be further discussed.

The fact that electro-magnetic energy may be transmitted along the inside of a single hollow conductor is perhaps not so familiar to the radio engineer, although the phenomena was demonstrated and studied theoretically in the 1890's, when radio communication was no more than a scientific plaything. Such a hollow tube transmission line is known as a " wave guide " and assumes great importance at wavelengths of a few centimetres since it has been found that with certain propagation modes the attenuation in a wave guide is much lower than in a coaxial cable. The absence of any insulating material in a wave guide is also a point to be noted, as in the coaxial line insulating spiders must be inserted at intervals, which still further increase the attenuation.

Wave Guide Fundamentals

Let us now see how the wave guide may be evolved from the well-known twin wire transmission line. A twin wire line may take the form of two flat parallel metal strips (Fig. 8A). Now decrease the wavelength of the energy applied to the end of the line until the width of the metal strips forming the line becomes an appreciable fraction of the wavelength. As the wavelength is reduced the intensity of the electric field across the width of the strips (which was more or less uniform at the higher wavelengths) will become more and more concentrated at the centre of the strips. A point will ultimately

be reached where the electric field will be very weak at the edges of the strips and very intense at the centre (Fig. 8B).

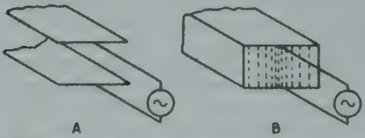


Fig. 8.
Derivation of wave guide from the conventional two-wire line.

When this condition has been attained then the edges of the strips may very well be connected together so as to form a long rectangular hollow box and energy will still be propagated along the line in spite of the short-circuiting sides of the box. The need for insulators has now been removed as no radio frequency voltages will appear on the outside of the wave guide if the metal used in the construction is thick enough. Note that there will be a " critical " wavelength above which the wave guide will not transmit energy. The cross-sectional dimensions of the guide must be of the order of a half-wavelength or more, *i.e.* the wave must " fit into " the guide if transmission is to take place.

Wave Guide Types

Wave guides may have rectangular, circular, or elliptical cross-sections. Various modes of propagation are indicated in Fig. 9. Where there is a component of electric force along the axis of the tube in the direction of propagation the waves are termed "E-waves." Similarly, if there is an axial component of magnetic force in the direction of propagation, the waves are termed "H-waves." There is an alternative nomenclature for the two types of wave, which is used by some American authors and which has been more widely adopted in recent literature on the subject. The types of wave are distinguished by referring to the transverse field in a cross-section of the guide, and the waves are called TE (transverse electric) and TM (transverse magnetic) waves. Thus the TE-wave which has an entirely transverse electric field and an axial component of magnetic force is equivalent to an H-wave, and a TM-wave having a magnetic vector in a plane at right angles to the direction of propagation and an axial component of electric force is equivalent to an E-wave.

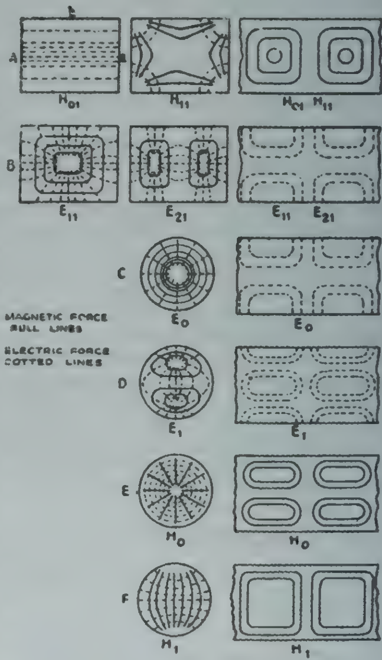


Fig. 9.
Modes in circular and rectangular wave guides.

In both systems the designating symbols may be given suffixes (which are either zero or positive integers). These suffixes indicate the mode which is being propagated in the tube. In the case of the rectangular-section guide, the subscripts *m*, *n*, denote the number of half-periods of electric field distribution along the two sides of the rectangle.

In Fig. 9A and B are shown four modes of excitation of a rectangular guide. The H_{01} mode is seen to be the simplest, having a sinusoidal electric field distribution across the side "a" only and the critical wavelength depends only on the length of the side "a." The other dimension "b" may be of any length for this mode. In order to transmit the longest possible wavelength for a given rectangular section the guide is usually operated with the transverse electric field across the longer sides of the rectangle. The other three modes illustrated are more complex. Each has its own critical wavelength. In general only the simpler modes of excitation are of importance, although when harmonics are present higher order modes may have to be considered.

When considering the field patterns in Fig. 9, it should be remembered that the lines of magnetic and electric force are everywhere at right angles—this being a fundamental characteristic of an electro-magnetic field.

In the case of the wave guide of circular section (Fig. 9 C, D, E, F) there is a similar series of modes and the various excitation modes are again distinguished by suffixes *m* and *n*. They are, however, interpreted slightly differently from the previous case. The subscript *m* shows the number of half-period variations of the angular component of the electric field when passing along a radius, and the subscript *n* indicates the number of full period variations of the electric field when passing around the circumference. If there is no variation in the distribution along either of these paths then the value of the subscript is zero. In circular section wave guides the subscript *m* is often omitted and is then understood to be equal to unity.

As in the case of the rectangular guide, there is also a simple mode for the cylindrical guide, the H_1 mode (Fig. 9F) in which the transverse electric field is more or less parallel to a diameter. This mode corresponds to the H_{01} mode for a rectangular guide. Other similar modes for rectangular and circular section wave guides may be seen by considering Fig. 9.

It is now apparent that the simpler types of cavity resonators which were discussed earlier are really short sections of wave guides closed at the ends by metal plugs just as in the case of the half-wave concentric resonator, although in the case of the wave guide resonator a half-wavelength measured in the guide is not necessarily equal to half of the free-space wavelength.

The modes of oscillation of such types of cavity resonators may also be classed as E- or H-waves and given distinguishing subscripts. The suffix will now consist of a number triplet, the first two numbers referring to the cross-section of the cavity as in a wave guide, and the third indicating the number of half-period variations in field intensity along a longitudinal section of the resonator.

Electrical Coupling to Wave Guides

Wave guides are usually excited by small probes carrying high frequency currents, inserted along a line of electric force, the probes being fed from a coaxial cable. The probes should run at right angles to the wall of the guide or to the end wall and must be so positioned to fall in a region of maximum electric field intensity for the desired mode. Where more than one probe is employed, attention must be paid to the phasing of the probe

currents. A movable metal piston (a " backing " piston) is usually situated behind the exciting rods, and its position is varied so as to obtain the maximum transfer of energy from the rods to the guide. Fig. 10 shows the location of the probes for the excitation of a guide in various modes. Energy may be transferred to a coaxial system at the far end of the guide by a similar arrangement of probes. As an alternative to the probes, coupling may be effected by introducing small loops into the guide, the loops being positioned in a region of maximum magnetic field intensity.

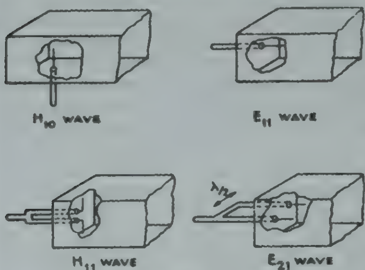


Fig. 10.
Probe coupling to wave guides.

Properties of Wave Guides.

In all wave guide systems there is an axial component of either magnetic or electric force. Hence, if boundary conditions are to be satisfied there will be a maximum wavelength above which no transmission of energy will take place. There is no axial component in a coaxial transmission line (Fig. 2) and so energy at any wavelength may be transmitted, but at very short wavelengths the diameter of the outer conductor must be kept small in relation to the wavelength to ensure that there is no spurious wave guide effect in the line.

The coaxial line is really a type of wave guide consisting of two coaxial surfaces, and if the diameter of the outer is large compared with the wavelength to be transmitted, the line may respond to a higher type of mode in which there are axial components of electric or magnetic force. For these higher modes the coaxial line will have a critical wavelength and will behave as a wave guide. Thus, if a tunable section of coaxial line of large diameter is used as a resonator, there may be several tuning positions for a given wavelength, because a half-wavelength for a " wave guide " mode will not correspond to a half-wavelength for the principal " tranmission line " mode. Now if the diameter is reduced so as to avoid the excitation of the line in a " wave guide " mode, at very short wavelengths the diameter of the inner line will become very small, giving rise to corona discharges between the conductors when high power is being transmitted along the line. This disadvantage, and the fact that the hollow pipe system has a lower attenuation and requires no insulators, makes the wave guide unequalled as a transmission line at wavelengths under 10 cms. (3,000 Mc/s.).

The velocity of propagation of energy in a coaxial line is approximately the same as the velocity of propagation in free space and hence the wavelength measured in a concentric line is the same as the wavelength in free space, but in a wave guide system the velocity of propagation in the guide is no longer equal to the velocity in free space, but depends on the frequency of the energy transmitted. At wavelengths much smaller than the critical wavelength the velocity in the guide is approximately equal to the velocity in free space, and the wavelength as measured in the tube approaches the value of the exciting wavelength. As the exciting wavelength approaches the critical value of the guide, the wavelength measured in the guide tends towards infinity.

All the above remarks refer to wave guides filled with air, but it may be noted that for a given tube cross-section the critical wavelength may be

increased by filling the guide with some suitable dielectric such as oil, but the losses in the guide will be very much increased and so the system is little used.

Obstacles inside a guide can be considered as lumped shunt impedances—provided they do not take up any appreciable length along the guide—and in particular forms are often used as matching devices. A projection into the guide from the wide side of a rectangular tube adds capacitance whilst one in the narrow side adds inductance. A load which appears inductive can, therefore, be matched by inserting a capacitive diaphragm as in Fig. 11A. Fig. 11B shows an inductive diaphragm whilst that in Fig. 11C is both inductive and capacitive and is, therefore, resonant, presenting no obstacle to the passage of the wave and not having a cut-off limitation due to the size of the hole. A completely blocked dead end to a guide acts as a short circuit, but an ordinary open end does not act as an open-circuit since it radiates and appears largely as a resistive load.

An important property of conventional transmission lines is the characteristic impedance—being the ratio of voltage to current at any point in an

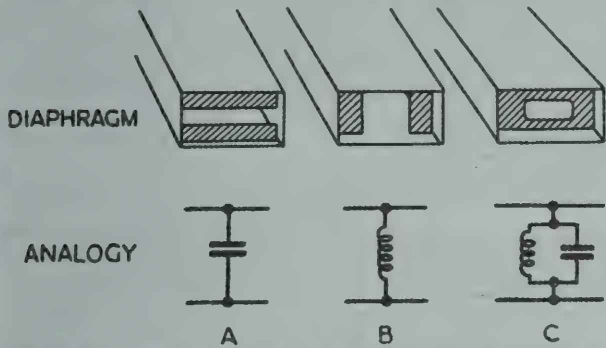


Fig. 11. Reactive diaphragms in wave guides.

infinite line. When a wave guide is considered, however, there is no return conductor and the usual concept of characteristic impedance must be revised. The " wave impedances " are defined as ratios of electric and magnetic field components. This involves the solution of the wave equations. For an E-wave in a uniform tube of any cross-section the wave impedance Z in the direction of the propagation is given by:—

$$Z = \sqrt{\mu/\varepsilon}\cdot(1 - [f_0/f]^2)^{\frac{1}{2}} = \sqrt{\mu/\varepsilon}\cdot(\lambda/\lambda_t)$$

and for H-waves:—

$$Z = \sqrt{\mu/\varepsilon}\cdot(1/[1 - (f_0/f)^2])^{\frac{1}{2}} = \sqrt{\mu/\varepsilon}\cdot(\lambda_t/\lambda)$$

where μ is the permeability of the dielectric
ε is the dielectric constant
f is the frequency transmitted
f_0 is the cut-off frequency of the wave guide

which may be calculated for the mode desired from the cross-sectional dimensions of the tube, and λ_t is the wavelength measured in the tube.

From the above equations it is seen that for E-waves the wave impedance increases from zero at the cut-off wavelength up to the value $\sqrt{\mu/\varepsilon}$. For H-waves the wave impedance decreases from infinity at the cut-off wavelength

down to the value $\sqrt{\mu/\varepsilon}$. For the standard rectangular wave guides the characteristic impedance lies between 350 and 500 at their operating frequency.

Relations between the critical wavelength and the cross-sectional dimensions of the guides for various modes are given in Table I. The figures apply for air-filled guides. In practice it has been found that wave guides need not be straight, but may be bent into any desired shape, without any appreciable amount of energy being lost at bends in the guide provided that the radius of the bends is not too small for the particular size of guide employed.

TABLE I

Tube Section	Mode	Critical Wavelength
Circular (Diam. $2a$) ..	E_0 E_1, H_0 H_1	$2{\cdot}61a$ $1{\cdot}64a$ $3{\cdot}41a$
Square (Side a) ..	E_{11}, H_{11} H_{01}* E_{21}	$1{\cdot}414a$ $2a$ $0{\cdot}894a$

Wave Guides in Practice

Right angle and other sharp bends may be made in a wave guide if certain rules are followed and the bends are of the form shown in Fig. 12A. Branches are also possible (Fig. 12B) but again they must be of certain form and certain dimensions are critical. Unless these design points are attended to the corners and branches give reflections due to mismatching and there is an appreciable loss of power.

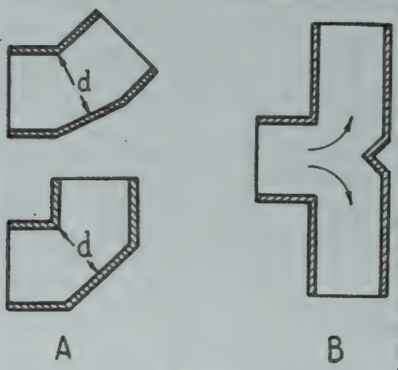


Fig. 12.
Wave guide bends and branch.

Wave guides have been standardised for the two important bands of 10 and 3 cms., being rectangular tubes of inside size 3 in. × 1 in., and supplied in 15 ft. lengths, and 1 in. × ½ in., supplied in 10 ft. lengths. The American standard sizes are slightly smaller than these (roughly by the thickness of the metal wall) and care must be taken in matching components to see which size is being used.

The usual material is copper as it has the lowest attenuation of common metals and can be easily bent and worked, but for short lengths of guide where the attenuation would, in any case, be negligible, brass can be used and is often more useful in experimental work. Due to the relatively short lengths in which wave guide is supplied and to the need to connect other components, wave guides have to be joined. The best way of doing this is by means of flanged joints as shown in Fig. 13A. The two flanges are held

* *Note that for this mode the guide may be of rectangular section as the critical wavelength depends only on the dimension of one side.*

together by a threaded ring similar to an ordinary pipe coupling, the ends of the guide butting together with the springy beryllium copper shim between them ensuring good contact. In particular the contact in the middle of the long sides must be good as a heavy current flows in these positions.

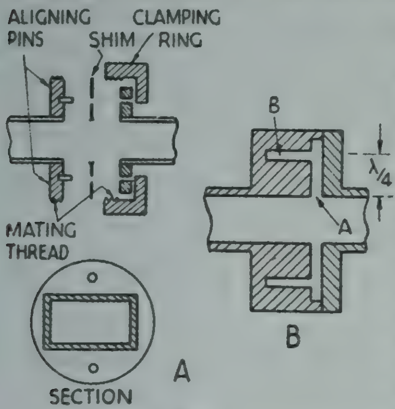


Fig. 13.
Wave guide couplings (A) plain, (B) choke.

An alternative method of jointing wave guides is what is termed the "choke joint," a representative type being shown in Fig. 13B. The edges of the two jointed guides are not in contact but are separated by an air gap A which is open-circuited by the resonant groove B. As the groove is a quarter-wavelength from the wave guide wall, the open-circuit appears as a short-circuit across the gap A at the junction of the ends of the guides, so that electrically there is no discontinuity in the wave guide. This type of joint is particularly useful in rotating joints such as are used in radar aerials. A choke joint although awkward to design and make is often better than a bad butt joint which can produce high losses.

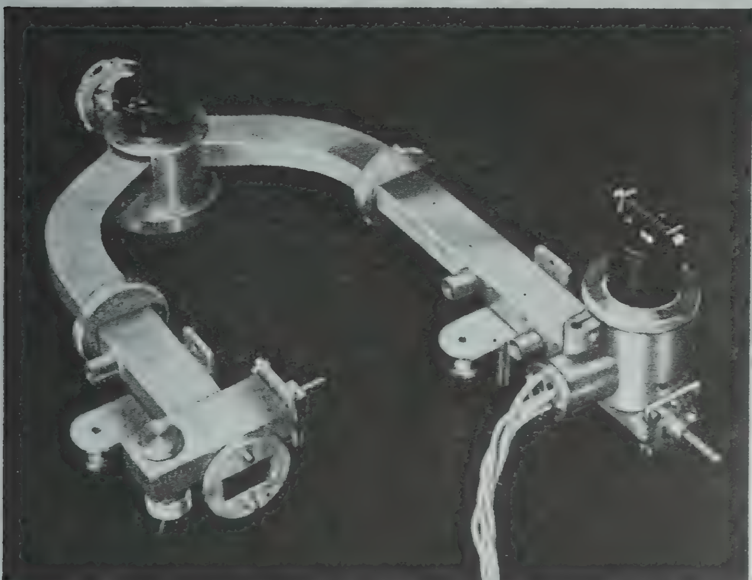


Fig. 14.
Typical 3-cms. wave guide assembly. *(Photo: courtesy of Marconi Instruments Ltd., St. Albans.)*

Switching of wave guides may be done by rotating plugs (somewhat similar to ordinary gas taps, although of infinitely more difficult design), or by systems of short-circuiting resonators across the switched arms. Many ingenious mechanisms have been tried and are described in the literature. Flexible wave guides find a use where it is desired to couple portable equipment and where there is considerable vibration preventing the use of rigid connections. Such guides are formed of rubber-bonded copper gauze tubes usually about 10 in. long with standard couplings at the ends of the section. Due to their high attenuations they are used only where absolutely necessary.

All these wave guide components, couplings, bends, branches, switches, rotating joints, matching sections, etc., are obtainable commercially (at least in the U.S.A.) and by connecting them together in the required manner, the whole of a R.F. section can be easily formed. By means of such a system of standardised parts with matching couplings and known characteristics the plumbing of a microwave set is made as simple to construct as the conventional L–C circuits of lower frequency equipment. Fig. 14 shows a typical assembly of 3 cms. wave guide and components. In the rear can be seen a three-way switch which allows the input coupling (extreme rear) to be switched to either the left or right-hand arm.

The left arm has an absorption attenuator, and a mixer-crystal mount formed of the crystal-housing in a short section of guide at right angles to the main wave guide. The absorption attenuator consists of a strip of carbonised insulating material placed longitudinally in the guide, which absorbs maximum power when in the centre of the guide and no power when flat against the narrow walls. The strip is moved backwards and forwards by a cam and spindle mechanism which is not shown. A backing piston in the crystal wave guide is used for adjustment.

The right-hand arm carries another attenuator and a reflector klystron oscillator which couples to the guide by a small probe. Another backing piston is used to adjust the klystron coupling. The wave guide screwed couplings which can be clearly seen in the illustration are of the direct type, using a metal shim to secure good contact.

CHAPTER 4. CIRCUIT ELEMENTS—VALVES

THE heart of any equipment is perhaps the valves and as would be expected the microwave frequencies have demanded special types. Research has proceeded along two lines; one to improve conventional valve types and adapt them to work at high frequencies and the other to produce entirely new types operating upon somewhat different principles. Before proceeding to discuss some of these new valves it may be of interest to examine the reasons for the failure of normal negative-grid triodes at very high frequencies.

Limitations of the Triode Oscillator

With a triode oscillator operating at normal frequencies it is customary to assume that the time required for an electron to travel from the cathode to the anode is negligible when compared with the length of time taken for one radio frequency cycle. At very high frequencies, however, this assumption no longer holds, and the electron transit-time becomes an appreciable fraction of the time taken for one period of oscillation. Fluctuations of the grid potential no longer cause an instantaneous variation of the anode current.

Suppose that the grid is biased negatively so that no grid current would flow at ordinary frequencies. Now if the grid were suddenly made more negative, the increase in grid voltage will prevent any more electrons being drawn from the cathode, but there will be a number of electrons which have already left the cathode and are travelling across the cathode-grid space. These electrons will be prevented from reaching the anode by the increased negative voltage on the grid, and on the anode side of the grid there will be a number of electrons in the grid-anode space which have just passed through the grid, and which are now suddenly repelled by the increased negative grid potential. Thus, the electron density on the cathode side of the grid will be greater than on the anode side, and hence electro-static charges will be induced on the grid structure and a momentary grid current will flow in spite of the fact that the grid has been made more negative. If the grid is now suddenly made more positive there will be a similar but opposite state of affairs, there being now more electrons receding towards the anode than there are electrons approaching the grid from the cathode. The effect is, that if a very high frequency voltage is impressed on the grid, an alternating current will flow in the grid circuit in spite of the fact that under static conditions no grid current would flow. This means that current is still flowing in the grid circuit when the maximum positive value of applied voltage is impressed on the grid, *i.e.* the input conductance is increased.

Ferris gives the formula for the input conductance as:—

$$g = K G_m f^2 \tau^2$$

where G_m is the mutual conductance of the valve

f is the operating frequency

τ is the electron transit-time for the cathode-grid space and K is a constant depending on the geometry of the valve and on the operating potentials.

The transit-time τ is constant for a given valve and for given anode and grid voltages so it is seen that the input conductance increases as the square of the frequency. This is the main factor which determines the maximum operating frequency of the negative grid triode oscillator.

The power loss in the anode is also increased by the transit-time effect due to the fact that the phase difference between the anode and grid voltages will now be less than 180°, and electrons will now be arriving at the anode when the anode potential has passed its minimum and is increasing, resulting in an increased power dissipation at the anode. This effect may be neutralised by introducing an impedance into the cathode lead of the oscillator thus adjusting the phases of the grid-cathode and anode-cathode voltages.

The maximum operating frequency of the triode oscillator is also limited by the fact that the valve itself contributes to the main oscillating circuit which is usually connected between the anode and grid. The physical dimensions of the external circuit are greatly reduced owing to the fact that there is a considerable amount of series inductance introduced into the oscillating circuit due to the length of electrode leads inside the wave envelope. Further, the main oscillating circuit is shunted by the internal inter-electrode capacitance of the valve.

Even if the lead inductance is minimised by eliminating leads through the pinch and employing some form of concentric lead-out for the grid and anode so that the grid and anode electrodes actually form part of the external transmission line circuit, the length of the circuit is still reduced owing to the loading effect of the inter-electrode capacitances. The loading capacitance is not simply the anode-grid inter-electrode capacitance, but has a larger

value, as it consists of the anode-grid capacitance in parallel with the anode-cathode and grid-cathode capacitances in series. (Fig. 15.)

Due to the fact that the grid input impedance of a valve with a load in the anode circuit is not the same as its impedance with zero anode load, the effective input capacitance is still further increased. This phenomenon is known as the Miller effect. A valve having a resistive anode load will have its grid-cathode capacitance C_{gk}, increased by the amount $(M + 1)\ C_{gp}$, where M is the voltage gain from grid to anode. Thus, neglecting other inter-electrode capacitances, the input capacitance will be:—

$$C_i = C_{gk} + (M + 1)\ C_{gp}.$$

A tuned circuit in the anode circuit of a valve will represent a resistive load and hence the input capacitance will be increased several times. If the anode circuit is detuned from resonance there will be a reactive component in the anode load and the effect will be that the input impedance is equivalent to that of a capacitance C_i and a resistance R in parallel. R will be negative for an inductive load, and positive for a capacitive load.

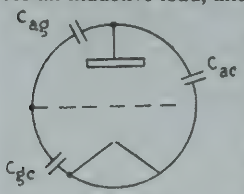


Fig. 15.
Triode inter-electrode capacitances.

It is seen, therefore, that if it is desired to increase the maximum operating frequency of the negative-grid triode oscillator, the following conditions must be observed:—

(1) The transit-time must be reduced either by increasing the anode voltage or by reducing the clearances between the valve electrodes. If the clearances are reduced, the physical dimensions must also be reduced in order to avoid increasing the inter-electrode capacitance. If this is done, the anode voltage cannot be increased due to the fact that the dissipation of the anode has been reduced.

(2) Some special form of electrode lead-out seal must be adopted to reduce the internal inductance of the valve. This may be done either by using a concentric lead-out or merely by using thick leads or a number of leads in parallel.

Small Clearance Valves

Acorn valves are perhaps the best known example of special pinch design. These valves have a normal frequency limit around 600 Mc/s., although a small usable output may be obtained at 800 Mc/s. if special care is taken in the design and layout of the oscillating circuit.

Another type is exemplified by the construction used in the popular midget television diodes. Triode valves are made on the same principle, being only slightly larger and with very small electrode clearances, and again perform up to about 800 Mc/s. Small negative grid triodes have been produced in the laboratory with a frequency limit of about 1,500 Mc/s., but the electrodes are so minute that only a few milliwatts of R.F. power are available. It seems feasible that triodes could be produced to oscillate at frequencies of the order of 3,000 Mc/s. without reducing the clearances below the values now used in acorns; but such valves could only find an application in receiver design and would be quite impractical for transmitting purposes.

Disc-seal Valves

Many valves intended for working in the lower frequency region of the microwaves use the parallel lead principle, the cathode, grid or anode or combinations of them having two, three or more different connections coming

out to separate pins on the base. It is found that a beneficial increase in input impedance is obtained especially in the case of high-slope valves where electrode lead effects are an important factor.

A logical development of this method of using leads in parallel is to make the lead a disc of metal, part of which forms the electrode of the valve. Recent advances in the technique of metal-to-glass seals have made this possible, and small triodes of the "lighthouse" and disc-seal types have made their appearance, permitting the use of a negative-grid oscillator at frequencies of the order of several thousand megacycles per second.

Fig. 16 shows the construction of the American lighthouse triode (A) and a British disc-seal triode (B). The electrode clearances in both these valves are similar to those in the acorn type previously mentioned. In the lighthouse valve the anode is formed from a metal cylinder, the flat end of which forms the anode proper. The cathode is also a metal cylinder, the emitting surface being the flat end. The grid consists of a metal disc with a centre hole over which has been welded a fine wire mesh. The glass envelope is thus in two parts separated by the grid disc. This valve is fitted with a standard international octal base and a concentric metal shell coupled to the cathode through an internal capacitance. The British disc-seal valve is similar in construction, but in this case the anode is pressed from a metal disc and the glass envelope is in three parts sealed to the electrode discs. The cathode and heater leads are intended to plug into a concentric type of socket.

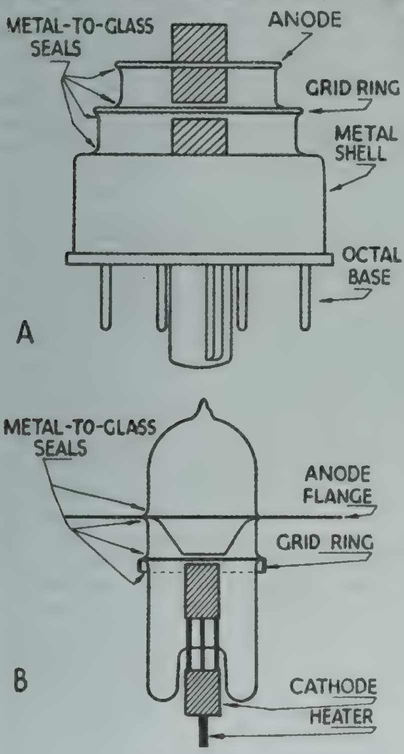


Fig. 16.
Disc-seal valves,
(A) American lighthouse type, (B) British type.

It will be seen that both these valves are designed for operation in concentric line circuits, and that the valve really forms part of the tuned circuit. For operation at very high frequencies it is usual to make the anode-grid circuit electrically a quarter-wave long and the grid-cathode circuit three-quarter-waves long. This, however, introduces difficulties in ganging the tuning plungers. Both the above types of valve are very suitable for use in receiver local oscillators as they will oscillate with an anode supply of 200–300 volts, thus obviating a separate high voltage supply which is necessary with other types of oscillator operating in this part of the spectrum. The use of a disc-seal valve is illustrated in a receiver circuit in Chapter 5 (Fig. 40).

Grounded (or Earthed) Grid Valves

A schematic circuit of a grounded-grid valve is shown in Fig. 17, where it will be seen that the grid of the valve is earthed and carries no signal voltage. The circuit has several advantages at U.H.F. over the ordinary grounded-cathode circuit due to the almost complete screening of the cathode from the anode and is used to a considerable extent in the band 30-5,000 Mc/s. As an amplifier, the circuit improves the noise-factor of receivers by some 8 db, a figure which results in a considerable extension of the range. The circuit has a low input impedance which makes it very suitable for following low output impedance stages, such as crystal mixers, and a low output capacity which is beneficial in wide-band amplifiers.

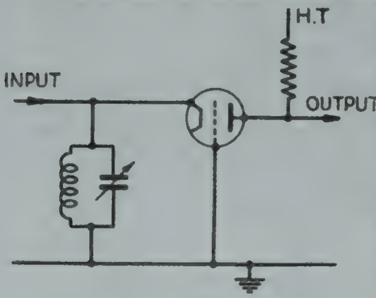


Fig. 17.
Circuit of grounded-grid amplifier.

The upper limit of operation of grounded-grid valves, whilst still offering advantages over other types, is about 2,000 Mc/s. for amplifiers and about 5,000 Mc/s. for oscillators. The majority of the valves employ the disc-seal technique just described, the grid connection being a copper disc sealed through the envelope with the grid filling a slot cut in the disc.

The Inductive Output Valve

There is, however, one type of valve developed by the R.C.A. which forms a link between the conventional type of triode oscillator and the newer types of oscillator employing velocity modulation, and it may be of interest to examine the operation of this inductive output valve before discussing the principles of velocity modulation. In this valve, electron transit-time effects are minimised by using an electron beam of high velocity, the resulting dissipation problems being solved by separating the functions of radio frequency output and current collecting electrodes. This allows the current collecting electrodes to be of large dimensions and thus powers of several hundred watts may be handled.

The output circuit consists of a quarter-wavelength of concentric transmission line, the inner conductor of which is hollow, and through which an electron beam passes. The outer conductor is continued as an extension of the inner, leaving a small gap between the ends of the two conductors. Consider the passage of a negatively charged body through the inside of the inner conductor (Fig. 18). A positive charge, equal in magnitude to the negative charge, will be induced on the inner wall of the inner conductor and will move along the inner wall as the negatively charged body passes along the tube. No charge is induced on the outside of the inner conductor. When the body is passing the gap between the ends of the conductors AB, part of the induced charge will appear on the end of the outer conductor (Fig. 18B) and when the body has passed the gap all the induced charges will be transferred to the extension of the outer conductor (Fig. 18C). For this to happen, a current will have to flow from A to B over the outer surface of the inner, and the inner surface of the outer conductor.

Now suppose the tank circuit to be excited at its resonant frequency. The electric field produced will be as shown in Fig. 19 and it is seen that it is

mainly radial and confined to the space between the two conductors. At the gap AB, however, there is a certain amount of fringing, and between the ends of the two conductors the electric field will have an axial component. The space inside the inner conductor and inside the extension of the outer is practically field-free and hence there will be no work done on a negatively charged body moving in the direction of the arrow until the charge enters the gap AB. If the charge enters the gap when the axial component of the electric field is in the direction B to A it will be retarded, part of its kinetic energy being given up to the tank circuit. If the charge should enter the gap when the field is reversed half a cycle later, it will be accelerated, the energy required to accelerate it being abstracted from the tank circuit.

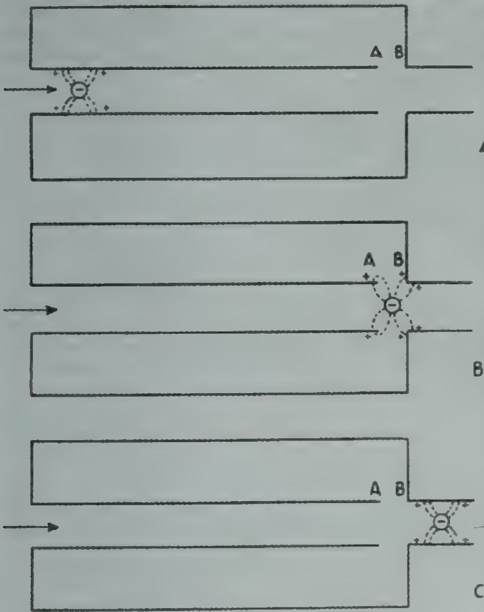


Fig. 18.

Passage of electron through a concentric λ/4 circuit.

It is seen, therefore, that in order to excite the output circuit it is only necessary to arrange that more electrons enter the gap during one half-cycle than during the following half-cycle. This can be conveniently done by introducing a grid structure so that, if the grid is excited at the resonant frequency of the output circuit, the electron beam will be broken up into a series of bunches of electrons travelling through the inside of the inner conductor of the output circuit.

The complete circuit of the inductive output valve is shown in Fig. 20. A conventional grid and cathode assembly is used to obtain pulses of electrons and the electron stream is concentrated at the start by means of the focusing electrode F. For further focusing, a solenoid is mounted concentrically

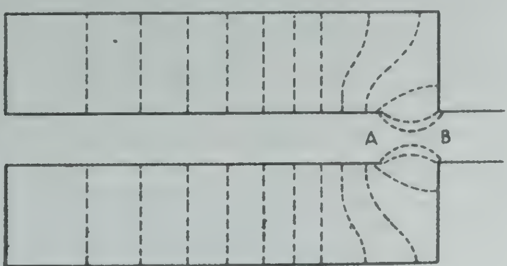


Fig. 19.

Configuration of electric field inside an oscillating λ/4 circuit.

with the tank circuit so as to produce an axial magnetic field.* Accelerating electrodes C and D are placed on either side of the gap AB, but are outside the alternating fields at the gap so that they do not constitute part of the output circuit. A current-collecting electrode E is mounted beyond the extension of the outer conductor of the output circuit in order to remove electrons which have given up part of their kinetic energy to the tank circuit during their passage across the gap AB.

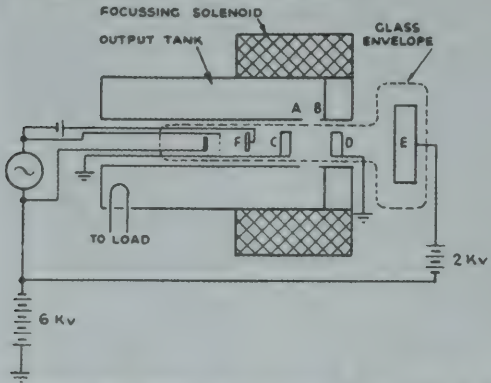


Fig. 20.
Inductive output tube.

When the excitation frequency of the grid is adjusted to the resonant-frequency of the output circuit there will be currents induced in the output circuit due to the passage of the bunches of electrons across the gap, and the phase of the high radio-frequency voltage which will be produced across the gap AB, will be such as to decelerate the electrons entering the gap during the half-cycle of maximum electron density, and thus energy will be transferred to the output circuit. During the next half-cycle when the electron beam entering the gap will have a minimum density, the radio frequency voltage will be such as to accelerate electrons in the beam, and thus energy will be transferred from the tank circuit to the electrons in the beam. But the electron density during this half-cycle will be very much less than during the previous half-cycle, so that the effect is that more energy is transferred to the output circuit than is abstracted from it. Energy may be drawn from the output circuit by means of a coupling loop inserted at the low impedance end.

This valve is intended mainly for use as a high-power radio frequency amplifier, but it may be used as an oscillator by providing a resonant circuit for the grid and coupling this to the output circuit.

So far, no mention has been made of the positive-grid triode (Barkhausen oscillator) nor of the magnetron oscillator, both of which make use of electron transit-time to maintain oscillation. Such oscillators have been employed for many years for the production of centimetre waves, but they really belong to the new class of velocity-modulated tubes and hence will be considered after the principles of velocity modulation have been discussed.

Velocity Modulation

We have now considered the more conventional type of microwave valve and one valve midway between ordinary types and really specialised high

* *See the* Radio Handbook Supplement, *Chapter 4, on Cathode-Ray Tube Focusing.*

frequency versions. It is proposed to pass on to a brief outline of those valves especially suitable for microwave frequencies and which, as mentioned earlier, use different principles of operation.

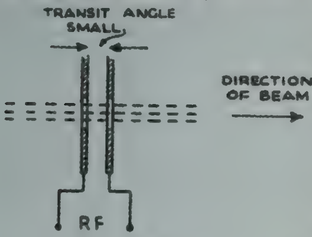


Fig. 21.
Velocity modulating grid.

Consider a beam of high velocity electrons passing through two grid electrodes AB, which may consist simply of the two plates with circular holes, across which a high radio frequency voltage is impressed (Fig. 21). It is assumed that the electron velocity is so high and the grids spaced so closely, that the time for an electron to cross the space between the two grids is only a small fraction of the time taken for one radio frequency cycle. This transit-time is usually expressed in degrees or radians and is known as the "transit angle," *i.e.* the angle described by the radio frequency voltage vector during a space of time equal to the transit-time. As there is a high radio frequency voltage across the two grids, A and B, there will be an alternating electric field between them which will be more or less parallel to the direction of motion of the electron beam. Suppose that when an electron is entering the inter-grid space the electric field is in the direction A to B. This electron will be accelerated and will leave the grid B with a velocity greater than the mean velocity of the electrons in the beam. Similarly, an electron entering the space when the direction of the field is from B to A, will be decelerated and will leave B at diminished velocity. Thus after leaving B the electrons in the beam will travel onwards with differing velocities, *i.e.* the beam has been modulated in velocity due to its passage between the grids. Note that the electron density of the beam on leaving B is the same as the density of the beam entering the gap; only the velocity of the electrons has been changed.

There is also a type of grid developed by Hahn and Metcalf which is suitable for velocity modulating a beam. This consists of two hollow tubes (Fig. 22) through which the electron beam passes. Suppose a radio frequency voltage to be impressed on the two tubes as shown. The transit angle between A and B, and between C and D is small. An electron entering the gap AB when the electric field across the gap is in the direction A to B will be accelerated during its passage. At this instant the field across CD will be in the reverse direction D to C. After passing B, at increased velocity, the electron enters the centre of the inner tube, which is field-free and so it

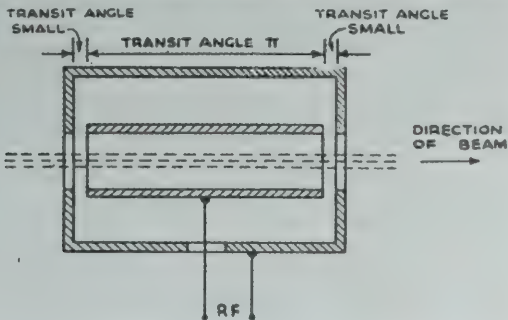


Fig. 22.
Hahn and Metcalf grid.

will proceed from B to C without further change of velocity. Now if the transit angle of the space BC is 180° (π radians), when the electron reaches C (half a cycle later) the field across CD will have reversed and will now be in the direction C to D which will again accelerate the electron. Similarly an electron which is decelerated at the gap AB will be decelerated at the gap CD and an electron entering the gap AB at an instant of zero field will pass through the grid and leave D with unchanged velocity. It will be seen that this double-acting type of grid is similar in action to the simpler type previously described, the electrons leaving D with differing velocities.

It now remains to be seen how the velocity modulation may be changed to amplitude modulation and the energy in the beam transferred to an output circuit. This may be done by three different methods: (1) deflecting the beam, (2) allowing the beam to drift in field-free space, (3) applying a retarding field to the beam.

Deflection Conversion

If an electron beam is allowed to pass through a magnetic field at right angles to its direction of motion, it will be deflected through an angle which is proportional to the strength of the field and inversely proportional to the velocity of the electrons in the beam. Hence if a beam containing electrons having different velocities is allowed to enter a steady transverse magnetic field, the slower-moving electrons will be deflected through a larger angle than the faster-moving ones. The main stream will be split up into two subsidiary streams one containing the fast-moving electrons and the other containing the slow-moving electrons. Thus the main velocity modulated beam has been converted into two smaller charge-density modulated beams from which power may be drawn.

A steady transverse electric field would also split a velocity modulated beam into two amplitude modulated beams. However, tubes using this method of conversion to amplitude modulation have proved rather difficult to adjust due to critical electrode potentials, and tubes using the two other methods of conversion have been found to be more satisfactory.

Drift Tube Conversion

If a velocity-modulated beam is allowed to pass through a field-free space (known as a " drift space "), the faster moving electrons will now begin to overtake the slower ones, so that at some point along the beam, the beam will consist of alternate regions of high and low electron density, *i.e.* the beam has become bunched and is now charge-density modulated at this point. The length of the drift space to produce bunching will depend on the magnitude of the radio frequency voltage existing across the electrode structure, used to produce the velocity modulation. Only a very small voltage is necessary to produce deep amplitude modulation, provided that the drift space is long enough, since the electrons will in time sort themselves out into bunches even if there is only slight velocity modulation when the beam leaves the modulating electrodes. However, very long drift spaces are not a practical proposition since space charges may be set up in them which have a deleterious effect on the bunching action.

It was seen that at some point in the drift space the beam was charge-density modulated, but the electrons in the bunches have different velocities and if the beam were allowed to continue to drift in field-free space they would unsort themselves after a certain time and again bunch themselves. Thus in a long velocity modulated beam passing through field-free space there will be several equally-spaced regions of charge-density modulation

along the beam, the maxima being spaced by a transit angle of 2π radians.

To abstract the energy from the beam it is only necessary to arrange for the output circuit to be placed at a point of maximum charge-density modulation. This output circuit may take the form of a concentric line circuit such as was used for the inductive output valve, in which case the mechanism for the abstraction of energy will be as previously described, or it may take the form of a Hahn and Metcalf grid (Fig. 22) coupled to a concentric circuit.

Consider a charge-density modulated beam entering the grid-structure shown in Fig. 22. It will be assumed that the grid is connected to a resonant circuit which is being excited. The cluster of electrons enters the gap when the field across AB is in the direction B to A. The field across the gap CD will be in the direction C to D at this instant. The bunch of electrons will be retarded as it passes through the gap AB and will give up energy to the oscillating circuit connected to the grid assembly. A half-cycle later it will reach the gap CD where the field has now reversed and is now in the direction D to C, so the bunch of electrons will undergo a further retardation and transfer more energy to the output circuit. When the fields are in such a direction as to accelerate (and hence abstract energy from the tank) there will be a region of minimum electron density entering the gap, so that there is more energy transferred to the output circuit than is abstracted from it. The electrons emerging from the grid structure may now be collected by means of a collector electrode, as is done in the inductive output valve.

Retarding Field Conversion

If a beam of electrons is allowed to approach a low potential electrode, the electrons will be retarded and brought to rest by the negative potential gradient, and will return along the beam. The point at which they are turned back depends on the velocity of the electron.

If a velocity modulated beam approaches a reflecting electrode at the potential of the cathode, the faster-moving electrons will have sufficient kinetic energy to reach the electrode and be collected. The slower-moving electrons, however, will be brought to rest before they reach the reflector and will return along the beam. This reflected electron stream is charge-density modulated, and may, therefore, supply radio frequency power to an electrode system placed in its path.

The Positive-Grid Triode or Barkhausen Oscillator

The positive-grid triode oscillator is one of the oldest devices for the generation of power at centimetre wavelengths and was accidentally discovered in 1919 by Barkhausen and Kurz when they were investigating the degree of vacuum in transmitting valves. In this oscillator a positive voltage is placed on the grid and a small negative voltage on the anode. There are several ways of connecting the resonant circuit to the valve, the most common being to connect it between the grid and anode.

Suppose the valve to be in oscillation and a radio frequency voltage to be impressed on the D.C. potential. All the electrons emitted by the cathode will be accelerated towards the grid and some will pass between the wires of the grid mesh. The impressed alternating potential on the grid will produce velocity modulation of the electron stream, which passes through the grid towards the anode, and as this is operated at zero or a small negative potential, the part of the electron stream which has the greatest velocity will reach the anode and be collected—thus an anode current will flow, this being the usual sign of oscillation in an oscillator of this type. Part of the stream is turned back before it reaches the anode and returns towards the grid. As

this stream is intermittent it is charge-density modulated and in passing through the grid it induces an alternating current in the grid structure as was described when considering the operation of the normal triode at very high frequencies. It now approaches the cathode and as its velocity has been reduced in its passage through the grid it will be again reflected by the cathode. It continues to pass to and fro through the grid until its velocity has been so reduced that it is collected by the grid. Thus if the external circuit is adjusted so that the time taken for one radio frequency cycle is equal to the electron transit-time between anode and cathode the oscillation will be maintained.

This type of oscillator is very inefficient as the electron stream is not focused and hence a large number of electrons will strike the grid without doing any useful work on the high frequency circuit. A development of this type of oscillator was used in the 17 cms. cross-Channel telephone link. In this case the valve used had a special helical grid with no supports running along it. The grid helix was designed to resonate at the working frequency and the Lecher wire output circuit was connected between the two ends of the grid.

Magnetron Oscillators

The magnetron valve first made its appearance in 1920, being designed for use on long waves. It consisted of a diode having a cylindrical anode and cathode, round which was wound a solenoid so as to produce a magnetic field coaxial with the cathode. A variation in the magnetic field would then control the electron stream flowing from cathode to anode. As the large inductance of the solenoid limits the use of this valve to very low frequencies it proved inferior to the normal triode and was ultimately abandoned.

However, around 1924 it was found that the magnetron could be made to produce oscillations of very high frequency by using a constant axial magnetic field and connecting a resonant circuit between the anode and cathode, or between the segments of a divided anode, this latter being the most usual method for the production of microwaves.

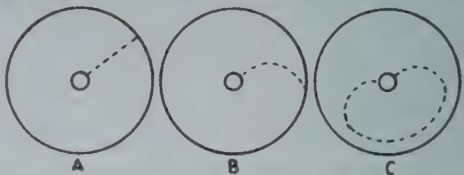


Fig. 23.
Magnetron electron paths,
(A) Zero magnetic field,
(B) Small field,
(C) Magnetic field at cut-off.

Consider a diode valve having a cylindrical anode and cathode, placed in a homogeneous magnetic field parallel to the cathode, and a D.C. potential applied between anode and cathode. When the field is zero the electron paths will be radial (Fig. 23A) and when the field is increased from zero the paths will tend to become curved (Fig. 23B) since the force acting on an electron moving in a magnetic field is always at right angles to its direction of motion. If the field is still further increased, a point will be found where the anode current falls to zero, the electrons now describing a circular path and returning to the cathode (Fig. 23C) after grazing the anode. Fig. 24 shows a magnetron characteristic.

In the case of the two-segment magnetron, which is one of the most important types for the production of microwaves, the anode is divided into segments by means of two longitudinal slots and an oscillatory circuit is connected between the two segments. The magnetic field is adjusted to the cut-off value. When the valve is in the non-oscillating state, both halves of the anode will be at the same potential and the electric field will be radial as

in Fig. 25A. The dotted curve shows the probable path of an electron under these conditions, and gives the direction of the force acting on the electron due to the electric field, and the direction of the force due to the axial magnetic field. Note that this is always at right angles to the direction of motion of the electron.

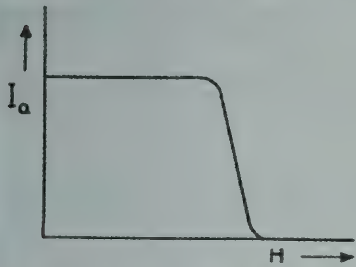
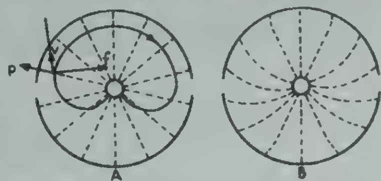


Fig. 24.
Magnetron characteristic.

Assume the tube is oscillating and that there is a small alternating potential between the two segments. The uniform radial electric field of Fig. 25A is now distorted by the application of an alternating tangential field, the distortion being strongest in the vicinity of the gaps (Fig. 25B). The electrons which pass in front of the gaps with the field will be accelerated, hence the

Fig. 25.
Split anode magnetron electron paths.
(A) Critical magnetic field and no r.f. on the segments. (B) Field configuration in oscillating split anode valve.



radius of their orbits will be increased, and the electrons will strike the segment with the higher potential (Fig. 26A) where their energy is dissipated as heat. Those electrons which pass the gaps, with an adverse tangential field, will be retarded and give up energy to the field and will continue on an orbit of smaller radius than the grazing orbit (Fig. 26B) towards the cathode. By the time the electron has returned to the vicinity of the cathode the field has reversed and the electron is again accelerated by the D.C. potential. It may repeat the same process during several consecutive cycles until its kinetic energy is exhausted and it can contribute no longer to the maintenance of oscillation.

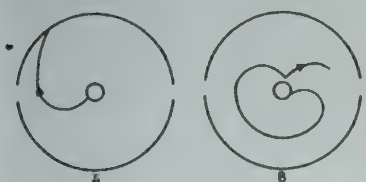


Fig. 26.
Oscillating split anode magnetron. Paths of electrons emitted when the upper plate is going positive. (A) Electron absorbing energy from output circuit. (B) Electron delivering energy to output circuit.

The frequency of oscillation of the tube is determined by the time taken for the electrons to complete their circular path from the cathode back to the vicinity of the cathode. It should be noted that the electrons which do not

contribute to the oscillating energy are automatically removed from the anode-cathode space. It has been found that an increased efficiency may be obtained if those electrons which have completed several cycles are removed. This may be accomplished by tilting the magnetic field a few degrees, so that the electrons describe a spiral path along the anode axis, or by fitting two end plates and applying an electric field in the same direction as the magnetic field. This will also give a spiral motion to the electrons, and when exhausted they will be collected on one of the end plates.

Very high frequencies may be attained by this mode of oscillation. Richter obtained a measurable output at 0·49 cms. (61,000 Mc/s.) using a tube of this type, although the efficiency was very low.

Cavity Magnetrons

A recent modification of the magnetron utilises tuned cavities; the essential difference being that the tuned output circuit is built into the valve itself. The anode assembly is machined from a solid block of copper, or built up from punched laminations, the tuned circuits taking the form of holes drilled round the internal periphery of the anode block and coupled to the internal cathode region by means of narrow slots (Fig. 27). The circuits may be coupled to each other at the ends of the anode block by means of copper straps joining equipotential points in adjacent circuits. It is sufficient to take the output from one of the tuned circuits by means of an output loop coupled into the cavity. This type of construction permits efficiencies of 65–70 per cent. to be obtained in places and 100 watts may be obtained at 30,000 Mc/s. (1 cm.) at an efficiency of the order of 50 per cent.

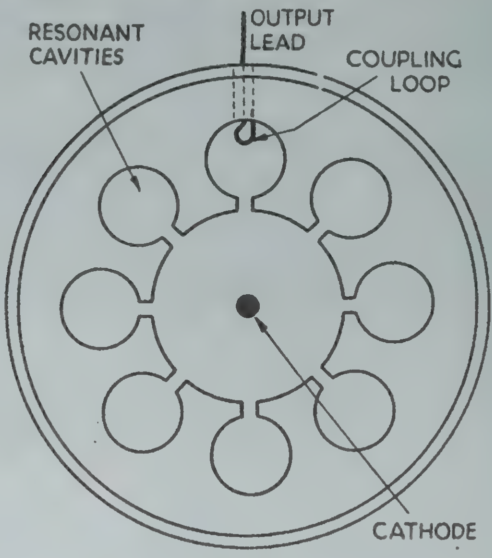


Fig. 27.
Anode block of cavity magnetron.

Such magnetrons are, of course, single frequency oscillators as it is difficult to alter the dimensions of the cavities in the evacuated space of the tube* whilst extremely accurate machining is necessary on the anode block to ensure that all the cavities have the same resonant frequency. Metal-to-glass seals are employed in bringing out the aerial lead and on each end of the anode block assembly.

By means of pulsing the magnetron with pulses of the order of $0 \cdot 1$–5μ sec. so that a high anode current flows for a very short time, enormously high output powers may be obtained. Figures of 25–50 kW. are quite common for small valves and particular valves have been made which give peak output of 4 megawatts (4,000 kW.). Of course, the pulse input has to be very large and is a difficult point to arrange in the design of pulsed magnetrons. A typical case would be a modulating pulse input of 1,150 kW. peak (244 v. at 43A) giving a R.F. output of 490 kW. at an efficiency of 47·5 per cent. The ratings of typical valves are given below:—

Wavelength, 50 cms.–3 cms. or lower (600–10,000 Mc/s.).
Anode Voltage, 2 kV.–30 kV.
Peak Anode Current, 5 A.–60 A. (pulsed).
Dissipation, 25 W.–1·5 kW.
Peak Power Output, 10 kW.–1,000 kW.
Efficiency, 20–60 per cent.
Magnetic Field, 1,000–5,000 gauss.

This valve represents one of the outstanding developments in British radar and was the valve which made possible high power centimetric radar. A representative 3 cms. cavity magnetron is illustrated in Fig. 28.

Travelling Wave Valves

The basic principle of the magnetron valve just described is the interaction between a travelling electric field (rotating around the anode block) with an electron stream travelling at about the same order of velocity and has been applied to a particular type of microwave amplifying valve. In the klystron the signal is used in the form of a stationary electric field applied to a

* *See reference to "preplumbing" in Chapter I.*

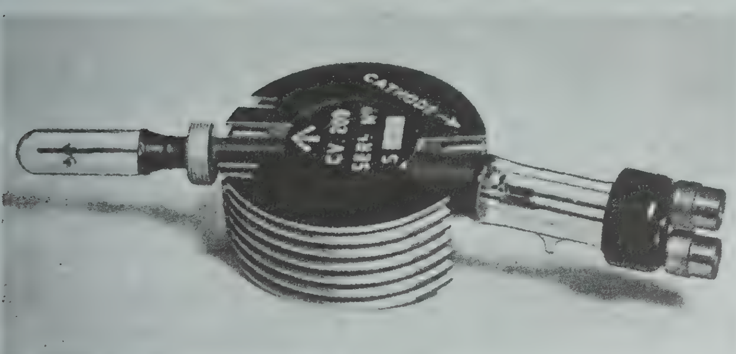


Fig. 28.

Typical 3 cms. cavity magnetron. (*Photo : courtesy of B. T.-H. Ltd., Rugby*).

rhumbatron but the travelling wave valve, as implied in the name, uses the signal in the form of a travelling electric field which exchanges energy with the electron stream of the valve. The electric field travels at the ordinary velocity of electric radiation (3×10^{10} cms./sec.) and has, therefore, to be slowed down to a reasonable velocity which can be given to an electron stream without recourse to excessive accelerating voltage.

This is done by making the wave travel down a helix of wire so that although the wave travels round and round the helix turns at roughly its normal velocity, its axial velocity along the length of the helix is reduced to about one-tenth of that value. If the electron beam is sent down the centre of the helix (Fig. 29), it will be density modulated—*i.e.* amplitude modulated—due to the interaction of the magnetic field of the helix with the electrons. This modulation " grows " along the length of the helix, roughly according to the square of the axial distance from the beginning of the helix. By suitable design this modulation can transfer energy to an external collector at the end of the helix and more energy can be extracted than was put in by the original signal, or in other words, the helix-electron beam mechanism is an amplifier. Some of the D.C. energy in the electron stream is being converted into A.C. energy in the wave.

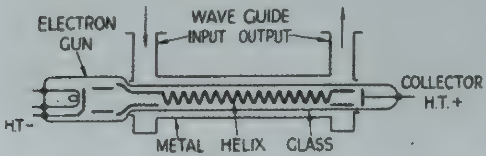


Fig. 29.
Travelling wave valve.

In its practical form the valve consists of the wire helix supported in a long glass envelope which fits through two wave guide stubs for input and output. At one end of the tube is a conventional electron gun assembly, at the other a collector electrode. An accelerating voltage of about 2 kV. is usual and a beam current of 1 mA. Power amplifications of up to 200 can be realised with the amazing bandwidth of 800 Mc/s. at an operating frequency of 4,000 Mc/s.

This very wide bandwidth is possible since the valve uses a completely untuned energy-exchange system, in marked distinction from magnetrons and klystrons which use a very sharply tuned resonant cavity. The bandwidth of a travelling wave valve is only limited by the coupling arrangements at the beginning and end of the helix. The valve opens up a new field of microwave endeavour and, in particular, allows the use of R.F. amplification prior to crystal mixers in receivers. In television and multichannel microwave telephony it should prove invaluable. A single valve could handle, say, 10,000 telephone channels or about 30–40 high definition television programmes at the same time.

The Klystron Oscillator

The klystron, one of the more successful of the new types of velocity-modulated oscillators, was developed mainly in the U.S.A. by R. B. and S. F. Varian. Fig. 30 shows the circuit. The device contains an electron gun similar to that in a cathode-ray tube, from which a narrow beam of electrons is projected along the axis of the tube and focused so that it will pass through the small apertures 1, 2, and 3, 4, across which the oscillating

circuits are connected. The transit angle of the two gaps 1, 2, and 3, 4, is assumed to be small. The resonant circuits are of the cavity type and consist of hollow toroidal chambers, known as "rhumbatrons." The oscillating currents in this type of circuit flow on the inner surface of the chamber and there will be a high radio-frequency voltage across the neck 1, 2.

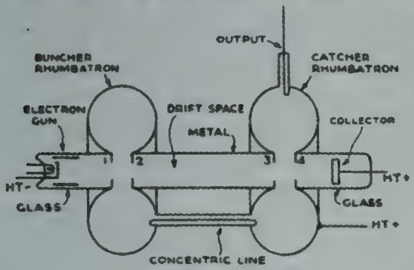


Fig. 30.
Klystron oscillator.

Suppose that the tube is oscillating, and ignore for the moment the length of concentric line connecting the two rhumbatrons. The electron beam from the cathode, accelerated by a D.C. potential of several thousand volts applied to the body of the tube, will pass through the aperture 1, 2, across which there is an R.F. potential. This will cause velocity-modulation of the beam in the manner already described. The velocity-modulated beam then enters the field-free space between the two rhumbatrons, known as the "drift space" where the faster electrons will begin to overtake the slower ones and the beam will become bunched. It should be noted that bunching in a drift tube really amounts to electron focusing in space and time. The distance between grid 2 and grid 3 is so arranged that the beam is bunched when it arrives at the latter, although it may, of course, be bunched and de-bunched several times prior to reaching this grid. If now we assume that when the bunches of electrons arrive at the second rhumbatron (the catcher) the electric field across the neck 3, 4, is adverse, the electrons will be decelerated in their passage and their lost energy is imparted to the field, and hence to the catcher rhumbatron. The electrons are retarded almost to zero velocity in their passage through the gap 3, 4, and are removed by a collector electrode which may be at the same potential as the body of the tube.

If oscillation is to be maintained it is thus seen that the beam must be bunched when it enters the catcher rhumbatron and that the phase of the R.F. voltage across the neck of the catcher must be such as to retard the bunches of electrons entering the aperture. This latter is easily arranged by feeding energy from the catcher to the buncher rhumbatron by means of a short length of concentric line, the length of which is varied so as to establish the correct phase relationship between the voltages across the two rhumbatron necks.

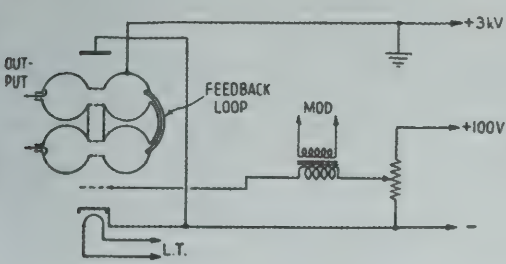


Fig. 31.
Power connections of klystron.

Tuning of the rhumbatrons may be done in two ways—mechanically deforming the cavity (one wall is often a corrugated diaphragm to allow this)—or by altering the volume of the cavity by screwing in metal plungers. Actual deformation of the cavity, results in tuning not so much by virtue of the volume change but mainly by altering the capacity between the entrance and exit grids of the resonator. The valve of Fig. 32A is tuned by this method; by compressing rings 1 and 2 for tuning the buncher, rings 2 and 3 for tuning the catcher. The reflector valve of Fig. 32B is tuned by the plunger method, having a large preset plunger for coarse tuning and a smaller one for variable fine control. The total range of tuning possible with any of these methods is only a few per cent., but in the case of a valve with external demountable resonators a greater change can be obtained by replacing the external parts of the resonators. Ganged tuning of klystrons is possible by mechanically coupling together the fine tuning controls, the coarse controls having been preset.

It should be noted that when the wavelengths of the rhumbatrons are altered, the accelerating potential on the body of the tube must be altered correspondingly in order to maintain the same transit angle in the drift tube. Note also that for a given wavelength there will be a number of accelerating potential values at which oscillation will be obtained—these values correspond to transit angles of $\pi/2$, $3\pi/2$, $5\pi/2$, etc. Maximum power output will be obtained with the maximum voltage on the body.

Electrical tuning of the resonators may also be done and in some cases has advantages over mechanical methods. By varying the accelerating voltage the transit-time and correspondingly the frequency is changed. Another method—used in valves employing high beam currents—is to vary the beam current, since by this means the dielectric constant of the beam varies, altering the capacity loading of the resonators. Both of these electrical methods have to be used with discretion since the frequency change is correlated with an output change.

R.F. is taken from the catcher rhumbatron by means of a short length of concentric line connected to a coupling loop, inserted in the rhumbatron, and which couples to the magnetic field round the axis. This valve, and the reflector version which is described in the next section has been an important one in the development of microwave use and was used a great deal in the early stages of centimetric radar research as a transmitting valve. It was later largely superseded by the cavity magnetron but has retained its usefulness as a local oscillator in receivers and measuring instruments.

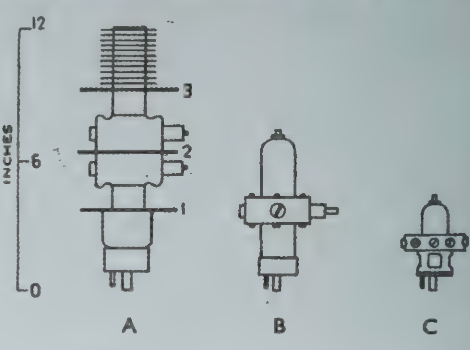


Fig. 32.

Types of klystrons. From left to right: A large transmitting type, a medium-sized reflector type, and a small receiving reflector type.

Modulation is usually effected by applying speech voltages to one of the focusing grids and thus varying the intensity of the electron beam passing through the necks of the rhumbatrons, although other methods are possible and in certain cases give superior results. Fig. 31 shows a typical circuit for the operation of a klystron valve; it will be noted that as is common in cathode ray tube work, the positive side of the H.T. supply is worked. This is because the external metalwork of the valve is connected to the rhumbatrons and would thus be normally at high potential. By suitable connections the klystron may be used as an amplifier, mixer or detector as well as an oscillator and thus has advantages over some other types of microwave valves. When the klystron is used as an amplifier the length of concentric line connecting the rhumbatrons is removed and the signal to be amplified is fed into the buncher rhumbatron. In the case of valves intended only for oscillator work, a permanent feedback line between buncher and catcher can be used, and such a line is frequently built-in the valve so that although it has two resonators there is only one connection—the output.

The sizes of klystrons vary from small tubes, no larger than, and plugging in like, ordinary receiving valves (used mainly in receivers and measuring instruments), to large ones with forced air or water cooling (used for transmitting) with accelerating voltages of perhaps 10 or 20 kV. Fig. 32 illustrates some of the sizes of double resonator and reflector type klystrons. The theoretical efficiency of a double-resonator valve varies between 58 per cent. for operation on the fundamental frequency to 24 per cent. for the 20th harmonic, whilst for reflector klystrons a figure of between 20 per cent. and 30 per cent. is possible. Ratings of typical valves vary between the following limits which illustrate the extremes in size:—

Wavelength, 50 cms.–3 cms. or lower (600–10,000 Mc/s.).

Accelerating Voltage, 250 V.–20 kV.

Beam Current, 30 mA.–0·25 A.

Dissipation, 5 W.–2·5 kW.

Power Output, 0·5 W.–200 W.

Efficiency, 10–15 per cent.

Amplification, 20 times.

Filament Rating, 4–6·3 V., 0·5–5 A.

The Reflection Type Oscillator and Reflector Klystron

This type of oscillator is rather similar in principle to the double-resonator klystron but employs only one resonant circuit or rhumbatron and is intended mainly for use in V.H.F. receivers where only a small amount of power is required. Figs. 33A and 33B show the circuits of two typical tubes, one employing a Hahn and Metcalf grid and a concentric line output circuit and the other using a rhumbatron.

The action of both these tubes is the same, depending on retarding field conversion. The electron beam is focused as in the klystron so as to pass

through the aperture in the Hahn and Metcalf grid (or rhumbatron) and is thus velocity-modulated on leaving the grid system or rhumbatron. It then approaches a reflector electrode which is at a potential near that of the

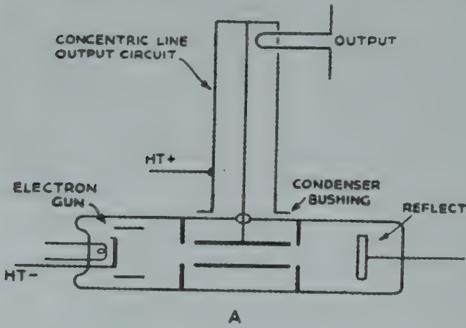


Fig. 33a.
Hahn and Metcalf tube.

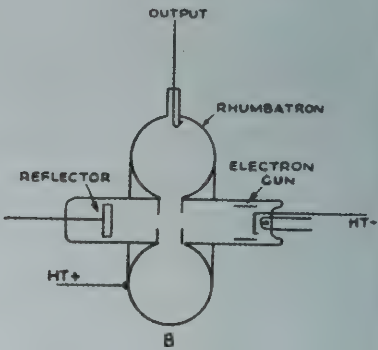


Fig. 33b.
Reflector klystron.

cathode, and may be either slightly positive or negative to the cathode. If it is at the cathode potential or slightly more positive, the velocity-modulated beam after leaving the modulating grid, enters a negative potential gradient and the part of the stream which possesses average, or more than average, kinetic energy will reach the electrode and give rise to reflector current. Part of the stream with kinetic energy below average will be retarded and turned back towards the cathode. This part of the stream is intermittent and is charge-density modulated.

Thus for oscillation to be maintained it is only necessary to arrange that the returning bunches of electrons are retarded during their second passage through the modulating grid or rhumbatron. If the reflector is operated with a potential negative to the cathode, no reflector current will flow and the entire beam will be reflected. In this case the returning beam is velocity-modulated and becomes bunched when it enters the modulating grid for the second time.

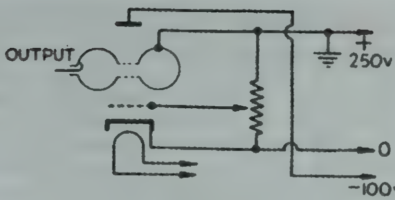


Fig. 34.
Power connections for small reflector klystron.

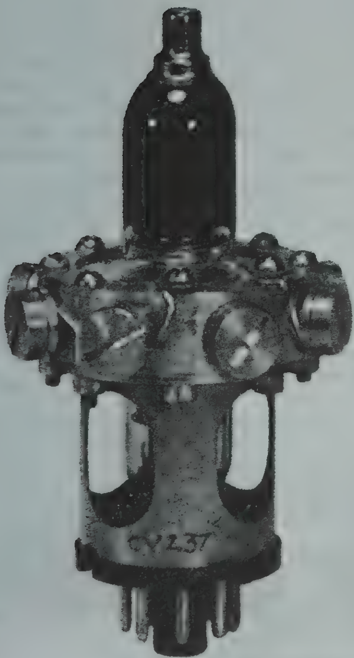


Fig. 35.
Small reflector klystron for receiver work.
(*Photo : courtesy of E. M. I. Ltd., Hayes, Middlesex.*)

The reflector klystron is a very convenient valve and although it is obvious that it can be used only as an oscillator, it possesses considerable superiority over the double resonator type for lower power oscillators. The reflector distance may be much shorter than the corresponding drift space in the double resonator valve due to the double transit of the electron beam and the low velocity of the beam over part of the travel. There is only one resonator to tune compared with the two of a double resonator valve which must be kept simultaneously and accurately in tune if any efficiency is to be realised. In other words, the buncher and catcher are always tuned to the same frequency as they are the same resonator. This advantage becomes peculiarly apparent when the trouble of initially setting-up a double valve has been realised.

Reflector valves have attained a wide popularity as local oscillators in V.H.F. receivers, due to this ease of tuning and their smallness and cheapness of construction. Fig. 34 shows typical power connections for a reflector valve, and Fig. 35 a small valve designed for receiver work, similar to that shown diagrammatically in Fig. 32C. An even smaller valve, with a spindle brought out for mechanical tuning, is illustrated in Fig. 14, Chapter 3, where the top may be seen projecting from the mounting on the wave guide.

Crystal Valves

The crystal detector, with its catswhisker and sensitive adjustment, has been known and used since the early days of wireless, although its use became

less and less with the advent of cheap and reliable thermionic valves. However, in 1941, the needs of microwave radar for a sensitive, compact and stable mixer or frequency converter led to the development of a vastly improved version of the old component. Early in the 1920's the use of silicon and germanium for crystal detectors had been reported and it was soon found that silicon was particularly suitable.

Fig. 36 shows the construction of a typical modern silicon crystal detector or crystal valve. A specially prepared flake of silicon is attached to the right-hand plug end and a contact of tungsten alloy wire is soldered to a plunger which slides through the left-hand head-end of the assembly. The enclosing tube is of low-loss ceramic material and, being silver-plated at the ends, is soldered to the two metal ends which form the terminals. A predetermined rectifying characteristic is obtained by adjusting the pressure of the contact on the crystal and the complete unit is then hermetically sealed. It will withstand considerable mechanical shock and a wide range of humidity and temperature without the performance being affected. The unit is slightly over ¾ in. in length and is supplied in three physical forms (small inserts in figure), for direct wiring into circuits, for plugging into special holders in wave guide assemblies, etc. (Fig. 14, Chapter 3, shows such an assembly), and a completely shielded type for terminating concentric lines.

Different electrical characteristics are available, suitable for use as mixer, second detector, modulator, etc. They can also be used for low frequency applications such as meter rectifiers and power work. Crystal valves are now almost universally used as mixers in microwave superheterodyne receivers and cannot easily be equalled in performance by other types of valve. Their great advantages are their small dimensions, small capacity, high stability and lack of necessity for any type of power supply. Similar valves, but employing a crystal of germanium are available and are said to have some advantages over the silicon type.

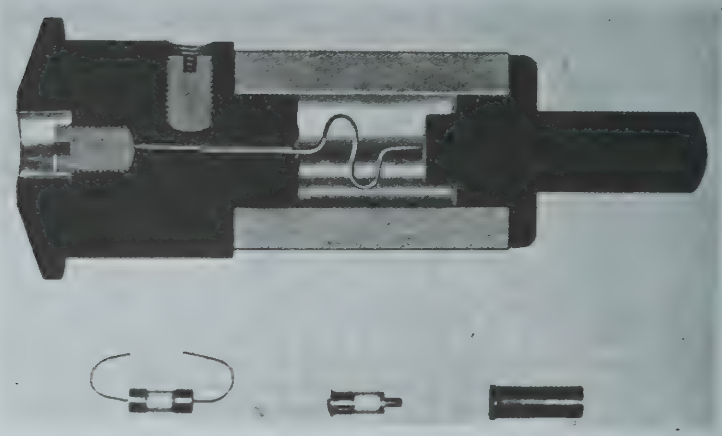


Fig. 36.

Construction of crystal valve. Inset—three types of mounting. *(Photo : courtesy of B. T.-H. Ltd., Rugby.)*

CHAPTER 5. TRANSMITTERS AND RECEIVERS

Transmitters

THE essential basis of microwave transmitters has already been covered in previous chapters, since the R.F. sections consist only of the special valves which usually have a built-in tuned circuit and the link to the aerial, either concentric line or wave guide—depending on the frequency. However, a few general considerations can be mentioned and two typical transmitters will be briefly explained.

Valves such as the magnetron which do not lend themselves easily to amplitude modulation may be used for communication purposes by employing some system of pulse modulation. One such system is variable pulse-width modulation. Pulses at a constant supersonic repetition rate are applied to the oscillator, so that R.F. energy is only generated during the duration of the applied pulse. The pulses are of constant amplitude and the application of speech signals to the modulator causes the width of the pulses to vary. Maximum modulation occurs when the space between successive pulses disappears; *i.e.* when C.W. conditions are reached. Such transmissions may be received on a conventional receiver since the pulse repetition rate is supersonic.

Various other systems of pulse modulation exist, including multiplex modulation in which several speech channels are carried on a system of interlaced pulses. These systems necessitate the use of an elaborate receiver to select the appropriate speech channels. The bandwidth required for all of the above systems is, of course, much greater than that required for pure amplitude modulation. Further methods include pulse-amplitude modulation, in which the amplitude of the pulses varies according to the modulation, and pulse-frequency modulation in which the repetition rate of the pulses is varied so that the space between them alters according to the modulation. Klystron oscillators may be easily frequency modulated by electrical means and some microwave communication systems employ such a method, or in other cases a klystron is used to amplify an already frequency modulated signal.

For non-communication purposes, such as radar and navigational aids, a pulsed R.F. output is used without any modulation. This has the advantage of producing a high effective peak power and a correspondingly greater range with relatively low dissipation in the output valves. In particular, magnetrons are favoured for this application and almost all microwave radar sets use a magnetron transmitter.

The most suitable type of valve for use in a high-power communications' transmitter appears to be the inductive output tube for wavelengths between 50 and 100 cms. and the klystron for wavelengths below 50 cms. Both these types are capable of outputs of several hundred watts and both may be used as high power radio frequency amplifiers, driven from a magnetron or other type of microwave oscillator.

The inductive output valve is easily modulated by applying speech voltages to the control grid, thus varying the intensity of the electron beam passing through the neck of the output circuit, at audio frequencies as well as at radio frequencies. As an alternative it should be possible to connect an absorption modulator circuit between the input grid of the valve and the output circuit of the driving oscillator. This modulator circuit may take the

form of a resonant circuit shunted by one or more valves acting as a variable load, audio voltages being supplied to the grids. This type of valve does not depend on velocity modulation for bunching the beam and hence the modulating potentials can be applied to the collector electrode as there is no drift space, the transit angle of which depends on the accelerating potential.

The disadvantage of the inductive output valve is that it requires a source of direct current for the focusing solenoid. There seems no reason, however, why a valve employing electrostatic focusing could not be produced at some future date.

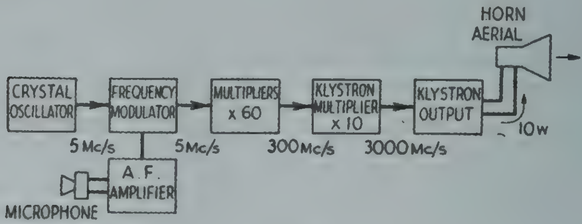


Fig. 37.
Block circuit of 3,000 Mc/s. f.m. transmitter.

For wavelengths under 50 cms. (600 Mc/s.) the klystron seems to be the best oscillator or amplifier for a communications transmitter. The simplest form of transmitter would be a klystron oscillator feeding the aerial directly, modulation being applied to a control grid situated next to the cathode and varying the intensity of the electron beam. Alternatively, modulation may be applied to one of the focusing electrodes. This latter method also varies the density of the beam passing through the orifice.

The block diagram of Fig. 37 shows a frequency modulated klystron transmitter for communication purposes. Up to the 300 Mc/s. stage the circuit is quite conventional, consisting of a master crystal oscillator, frequency modulator (Armstrong system) and multiplier stages. A klystron multiplier then increases the frequency to 3,000 Mc/s. and a final klystron amplifier delivers 10 W. to the aerial. A multiplication of ten in the first klystron is quite feasible, since the nature of velocity modulation allows this at frequencies where the rhumbatrons are of reasonable size. The final rhumbatron of this valve and the two in the power amplifier are fixed-tuned at 3,000 Mc/s. and no difficulty is experienced in lining them up. In contrast

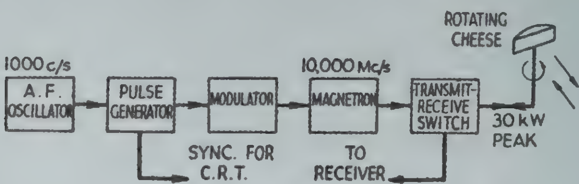


Fig. 38.
Block circuit of 10,000 Mc/s. magnetron transmitter for marine radar equipment.

to this transmitter for communication, Fig. 38 shows a pulsed magnetron transmitter for radar use. Its much greater simplicity will be noted. A 1,000 c/s. pulse generator giving $0 \cdot 2\mu$ sec. pulses feeds a hydrogen thyratron modulator which in turn pulses the 10,000 Mc/s. cavity magnetron. From the magnetron the R.F. is led to the aerial (in this case a rotating cheese) by a standard 3 cms. wave guide. Although the mean power output of the magnetron is only a few watts, the use of a pulse output raises the peak output power to some 30 kilowatts.

Super-Regenerative Receivers

The simplest type of microwave receiver is the super-regenerator, good sensitivity being obtained with the minimum number of valves. For wavelengths down to about 40 cms. (750 Mc/s.) acorn triodes may be used as oscillators either single-ended or push-pull. Quenching is brought about by allowing the grid circuit to block the oscillation periodically. This self-quenching is not usually possible near the oscillation limit of the valves, and an auxiliary interruption frequency oscillator may have to be resorted to if the receiver is to be tuned down to the minimum wavelength attainable.

Under 40 cms. oscillators of the reflection type (Fig. 33, Chapter 4) can be used in conjunction with a quench oscillator, the aerial being connected to the tank circuit or rhumbatron, and the quench voltage injected into either the body or the reflector electrode. The valve shown in Fig. 33A was designed for operation in a receiver of this type. Good results are obtained with such simple receivers, but they suffer from the same disadvantages which are exhibited at higher wavelengths, namely, high unwanted radiation (causing interference to neighbouring receivers over a fairly wide range of wavelengths), and a very high background noise in the absence of a received carrier. Their simplicity, however, makes them well worthy of consideration if cost is to be a deciding factor in design.

Superheterodyne Receivers

As at higher wavelengths, the superheterodyne is the most satisfactory type of receiver. It has good sensitivity and selectivity and a good signal-to-noise ratio. Furthermore, unwanted radiation from the oscillator is not so troublesome as it is in the case of the super-regenerative receiver. It has a great advantage over all other types as regards wavelength coverage, since the second and third harmonics of the oscillator can be used to beat with the signal, thus allowing reception down to very short wavelengths using an acorn-type oscillator operating on the long centimetre waves.

The main difference between a microwave superheterodyne and one designed for use on normal wavelengths lies in the mixer circuit. If it is only desired to receive in the vicinity of 100 cms. (300 Mc/s.) then a pentode type acorn can be used as a mixer and normal U.H.F. superheterodyne practice followed. Below this wavelength, diodes or crystal valves are used. Any non-linear device which has oscillations at two frequencies impressed upon it will produce beat notes, hence any diode (or triode) may be used as a frequency converter.

In Fig. 39A is shown the radio frequency section of a 40 cms. (750 Mc/s.) superheterodyne used in an aeroplane blind-landing equipment. The third harmonic from an acorn-type oscillator is injected into the diode mixer circuit where it beats with a signal from the aerial, for form an I.F. signal which is

applied to a high-gain amplifier. In the simplified schematic diagram in Fig. 39B the input circuit A is tuned to the signal frequency, and B to the oscillator fundamental. Circuit A is effectively a length of capacity-loaded concentric line while B is a one-turn coil consisting of a length of copper tube resonated by a small variable condenser and tuned to approximately three times the signal wavelength. The signal from the input circuit A is applied to the anode of the diode by means of a small condenser (built into the central conductor of the input circuit in Fig. A). As the coil of circuit B has appreciable inductance at the signal frequency, its tuning condenser acts as a bypass condenser for signal frequencies. The oscillator voltage (at 120 cms.) is introduced into circuit B by means of a small coupling loop tapped on to the coil. The diode anode coupling condenser is large enough to act as a bypass condenser to the oscillator voltages and the signal frequency circuit A offers but little impedance at three times the resonant wavelength, so that for all practical purposes the diode is connected across the tuned circuit B. Since the diode is a non-linear device, harmonics will be generated in the diode, and the third harmonic of the oscillator will mix with the signal from the input circuit to produce a beat at the intermediate frequency (in this case 10 Mc/s.). The anode condenser of the diode presents a high impedance to the I.F. voltages generated and hence the I.F. input circuit is connected across it.

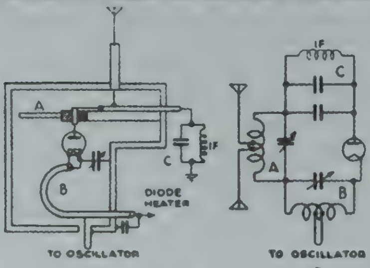


Fig. 39.

Diode mixer circuit of 750 Mc/s. superheterodyne receiver.

The diode mixer is obviously incapable of providing any gain and has also the disadvantage that the I.F. oscillations are present in the same circuit as the signal and local oscillator voltages, and hence the I.F. may beat with the local oscillator to produce unwanted frequencies. Also, as the diode introduces considerable damping into the input circuit, the Q of the signal circuit will be very much reduced, resulting in a low image-rejection ratio, and as oscillator and oscillator harmonic voltages are present in the input circuit there will be a fair amount of radiation from the aerial. This is not so serious as in the case of the super-regenerative receiver, because the radiated signal is pure C.W. In one paper, conversion " gains " of 0·6 to 0·8 are quoted for small " television " type diodes which have been used as mixers down to 50 cms. (600 Mc/s.).

The intermediate frequency amplifiers used in microwave superheterodynes resemble those used in television receivers. The frequency must be high enough to obtain a fairly wide band-width so that oscillator drift does not become objectional, and to allow unintentionally frequency modulated transmissions to be received. The intermediate frequency transformers may be loaded with resistors in order to achieve a wide band-pass (of the order of 0·25 to 0·5 Mc/s.). Such a wide-band I.F. amplifier should make the tuning of the receiver equal to that of a broadcast receiver. Three or four stages of I.F. amplification are usual as the damping resistors connected across the coils reduce the stage gain.

As the wavelength of the signal circuit is decreased the frequency of the intermediate frequency amplifier must be increased as a greater bandwidth is required for stability reasons. The following empirical formula has been

quoted by Dudley for determining the value of the intermediate frequency to be used in terms of the bandwidth and image ratio.

$$f = \tfrac{1}{4}\,\Delta f\,E_s/E_i$$

where Δf is the bandwidth of the receiver

E_s is the signal frequency voltage

E_i is the image frequency voltage.

There is another type of I.F. amplifier which may prove to be useful in the microwave spectrum—the resistance-coupled video amplifier. Such an amplifier will provide a good overall gain from the high audio frequencies up to 120 kc/s., or higher. With a set of this type the construction is very much simplified. The oscillator can be used as a combined oscillator-mixer as the signal and local oscillator are so close in frequency and the I.F. amplifier requires no lining-up. If desired, the I.F. response may be peaked by including a choke in the anode circuit of the first I.F. stage. Using a set of this type the signal and image occur very close together on the tuning dial. Radiation may become rather excessive since the aerial is coupled directly to the local oscillator.

A typical receiver for frequencies in the region of 4,000 Mc/s. is shown in Fig. 40. A disc-seal triode oscillator is used in a concentric line circuit, the anode-grid circuit being a quarter-wave and the grid-cathode circuit being three-quarter-waves long, both being tuned by movable metal plungers fitting in the valves. The third harmonic of the oscillator (1,500 Mc/s.) is used. The mixer consists of a section of rectangular wave guide operating on the H_{01} mode, the guide being closed at both ends and tuned by a movable shorting plunger at one end. The signal frequency is 4,500 Mc/s. Signals from the aerial and the oscillator third harmonic output are injected into the tuned wave guide circuit by means of probes projecting into the guide. The concentric line to the first I.F. circuit is terminated in a small crystal detector capsule which is connected across the cavity parallel to the electric field. The I.F. signal generated in the crystal detector is then fed to a high-gain 45 Mc/s. amplifier. It will be realised that the tuning of such an oscillator-mixer becomes rather complicated mechanically as the three tuning plungers move at different speeds and are usually rack-and-pinion driven.

Another microwave receiver, this time for 10,000 Mc/s. is shown in Fig. 41.

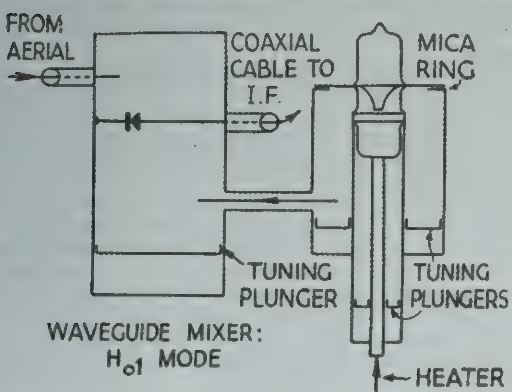


Fig. 40.
Wave guide mixer of 4,000 Mc/s. receiver.

The signal is fed straight into a crystal mixer, together with a local oscillation derived from a quartz crystal oscillator and multiplier, and klystron multiplier system and the I.F. is 45 Mc/s. Where variable frequency operation is

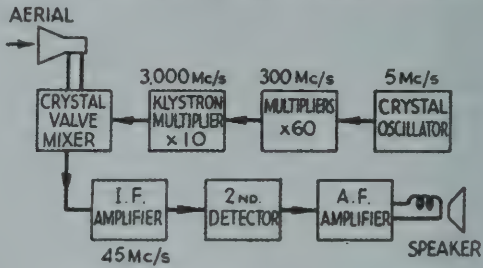


Fig. 41.
Block circuit of 10,000 Mc/s. receiver.

needed the crystal control is, of course, not used and a simple klystron oscillator can then be employed, with A.F.C. derived from the I.F. amplifier if desired. In another type of construction, the signal is led down a wave guide on to which is directly mounted the crystal mixer and klystron oscillator. Parts of the assembly shown in Fig. 14, Chapter 3, are representative of this type of receiver, so that the external circuits of the set consist only of the I.F. and A.F. amplifiers.

CHAPTER 6. AERIALS

Dipole Aerials

THE small dimensions of dipoles in the microwave wave-band makes the construction of highly directive radiating systems a practical proposition. Full use is made of reflecting surfaces and aerial arrays begin to resemble light projectors. For the longer waves (100 to 50 cms.) aerials may be similar to those used at wavelengths around 5 metres, employing reflector and director waves.

In Fig. 42 several types of directive arrays using reflecting surfaces are shown. The "bill-board" array (Fig. 42A) consists of a number of stacked dipoles erected at either $\lambda/4$ or $3\lambda/4$ from a reflecting metal sheet. This reflector sheet may consist of fine-mesh wire netting in order to reduce the wind resistance. Simpler types of highly directive aerials using only one dipole are shown in Fig. 42B, C and D.

Paraboloid and parabolic reflecting surfaces with a dipole at the focus give the sharpest beams, in fact a beam like a searchlight can be obtained with a paraboloid provided that the aperture is of the order of 10 or 20 wavelengths. This rather limits the use of the paraboloid to the very short centimetre waves if the size of the array is a limiting factor. In the 17 cm. cross-Channel telephone link paraboloids were used, having a dipole at the focus, and a hemispherical reflector in front of the dipole, so that all the radiation was thrown back into the paraboloid and an extremely sharp beam obtained. A gain of 33 db was realised with this set-up.

Perhaps the simplest type of directive array for use on centimetre waves is the " square-corner " beam (Fig. 42D). This consists of two metal reflecting sheets arranged at an angle and the dipole situated at the bi-section of the angle. For amateur purposes the reflector sheets may be replaced by a number of wires, one half-wave long, arranged in two intersecting planes, thus making the system very light and reducing the wind resistance. The beam obtained will not be so sharp as that given by the parabola or paraboloid, but the array has the very great advantage that it requires no adjustment when once set up.

Horns

One of the most interesting types of radiator is the electro-magnetic horn which consists of a wave guide flared out at one end. This type of radiator can be used only at wavelengths where wave guide systems are of practical dimensions. If oscillations are set up in a wave guide they travel along it, and if the remote end is left open they are radiated into free space. The radiation pattern depends on the cross-section of the wave guide, a large cross-section (compared with the wavelength) giving a more directive beam. This is due to the fact that the waves undergo diffraction when passing from the guide into free space. The amount of diffraction depends on the size and shape of the aperture through which the waves pass. This phenomenon is analogous to the passage of a beam of light through a narrow slit of width comparable to the wavelength of light. A sharp beam can be obtained from a simple open-ended wave guide but the dimensions must be several times the wavelength employed. That means that the guide must be operated well below its critical wavelength where the wave impedance is approaching the value $\sqrt{\mu/\varepsilon}$ (see Chapter 3). This value is that obtained for waves in an unbounded medium and is termed the " intrinsic impedance " of the medium or in the case under consideration, the intrinsic impedance of free space.

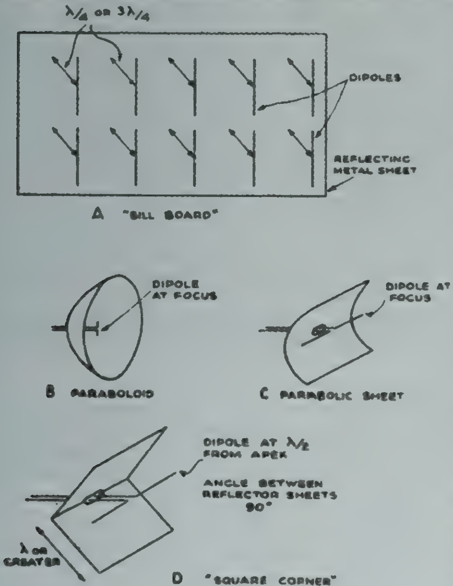


Fig. 42.
Directive microwave aerial systems.

In order to make the wave guide of economical dimensions it must be operated near its critical wavelength and if a sharp beam is desired the wave impedance must be matched to that of free-space. This matching is done by flaring the open end of the tube in one or more directions so that the aperture through which the waves pass into free space is of the order of 10 to 12 wavelengths.

In Fig. 43 is shown a short length of wave guide with the end flared out in one direction giving a beam which is sharp in the plane of the parallel sides. The wave guide portion of the horn, which may be of any length, has H-waves set up in it. As the waves pass into the horn section their velocity varies along the length of the horn. It will be remembered that the velocity of propagation in a guide operating near its critical wavelength is greater than the free space value. Thus the wavelength of the waves will change along the length of the horn. Fig. 43 also shows this field distribution in the horn section.

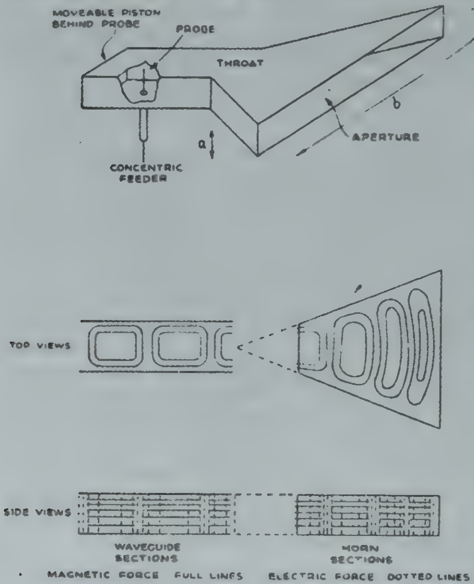


Fig. 43.
Horn aerials.

The horn radiator offers several advantages over other types. It requires no insulators and owing to the fact that the feed is remote, spurious lobes are absent in the radiation pattern. It is easy to set up, the only adjustment being the backing piston behind the exciting probe in the wave guide section, and it may be operated over a wide range of wavelengths, of the order of two to one.

When considering cavity resonators, wave guides, and horn radiators it will be seen that there is a certain amount of similarity to acoustic resonators and radiators. This similarity will be observed in any system where the length of the waves used is of the same order as the cross-section of the boundary walls.

Cheese Aerials

The " cheese " aerial has been used to a great extent in marine radar work since it produces a fan-shaped beam that is sharp only in the horizontal plane. By this means, the effect of the ship's rolling and pitching is reduced and the broad vertical beam allows both surface and air targets to be detected at the same time. It consists of a metallic mirror (Fig. 44A) in the form of a paraboloid section fed from a small horn at its focus. The long horizontal dimension and the short vertical dimension of the cheese produce the desired beam characteristic. The top and bottom of the reflector can be closed, as shown, or left open, depending on the mechanical construction preferred. For radar uses the complete cheese and horn rotate so that the whole horizon is scanned. For ground use a half-horn as Fig. 44B is sometimes used, and has some advantages, one of which is that the working of the aerial is almost independent of frequency.

Slot Aerials

A rather unusual type of aerial is the slot aerial. It has been found that a radiating slot in a metal sheet is equivalent to a radiating dipole; the principle is, in fact, a radio application of a well-known optical law. Slots are usually used in arrays and are cut in the sides of wave guides so that they either intercept currents flowing in the walls of the guide or intercept no current but have current drawn across them by coupling loops or probes inside. They are usually about a half-wave length long and are distributed in a definite pattern depending on the wavelength. Impedance values of about 485 ohms are normal; the selectivity of the slot being decreased by making it wider, just as an ordinary dipole is " broad-banded " by fattening the two arms.

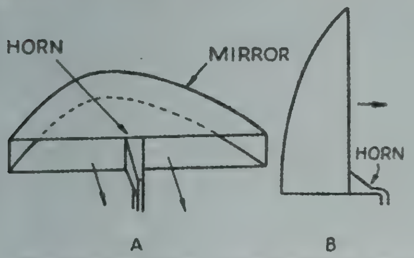


Fig. 44.
Cheese aerials.
(A) Rotating marine type.
(B) Vertical half-cheese.

The simplest form of slot aerial is the open end of a rectangular H_1 wave guide, although in this case the slot has a very low Q and is badly mismatched. A future use of slot aerials is on high-speed jet-propelled aircraft where an array of slots can be cut in the metal skin of the plane, the spaces being plugged with a dielectric material. By this means a perfectly smooth, dragless aerial can be realised.

It will be seen that the types of microwave aerial are many and various and although only a selection has been covered in this review, most of the representative ones have been indicated. In ever-increasing measure, such aerials take on the familiar form of optical systems, and indeed a standard design technique is to start from an optical analogy.

CHAPTER 7. MEASUREMENTS

MEASUREMENTS at microwave frequencies present unique difficulties and because of the complexity and ramifications of the subject, it is proposed only to touch upon two of the most important aspects of this important branch of microwave technique, namely, measurements of wavelength and power.

Wavelength is mentioned specifically, and not frequency, because at microwave frequencies it is nearly always wavelength that is measured. Although, in most cases, there may be assumed to be the usual relationship between frequency and wavelength, it must always be remembered that the two are quite different, one having the dimensions of the reciprocal of time and the other obviously of length. Converting from one to the other—using the standard wave velocity of 3×10^{10} cm/sec.—can only be achieved when it is definitely known that this value does obtain. Due to the shortness of the waves and physical size of the apparatus employed, it is usually far more convenient to measure wavelength.

Next to knowing the wavelength of a signal, we want to know the amount of the signal and the measurement of power satisfies this requirement. The direct measurement of either voltage or current at microwave frequencies is difficult, but power measurement is relatively easy and is normally adopted.

Wavelength

In the neighbourhood of 100 cms., wavelengths may be measured fairly accurately by measuring the standing waves on Lecher wires as is done at higher wavelengths. As the wavelength is decreased, the spacing between the rods or wires must also be decreased due to the increasing inductance of the shorting link, and a limiting point is reached. A better form of wavemeter consists of an odd number of quarter-wavelengths of concentric line with the inner conductor adjustable in length by means of a micrometer lead screw, and the conductors shorted at one end. Resonance is indicated by means of a crystal detector (or diode) and microammeter coupled to the low impedance end of the line. The input to the wavemeter is also connected at this point. The distances between maxima as measured on the micrometer scale give half-wavelengths directly.

Wavemeters of this type may be employed down to wavelengths of the order of 10 cms. with fairly good accuracy provided that the diameter of the outer line is sufficiently reduced, in order to avoid " false " maxima due to wave-guide resonances in the line.

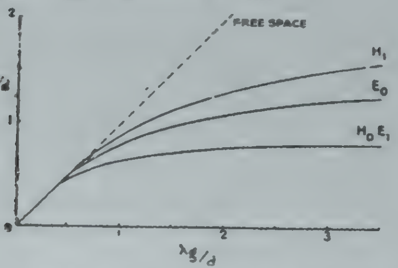


Fig. 45.
Relation between the free space wavelength and the wavelength measured in a circular wave

It will be remembered that in a wave guide operating near its critical wavelength, the wavelength measured in the guide may be much greater than the free space value due to wave velocity varying, as suggested above. This property may be used to measure very short wavelengths with much greater accuracy than can be obtained using a concentric line wavemeter. The relation between λg (the wavelength in the guide) and the free space wavelength, may be calculated from the field equations. A conversion curve for a circular section guide is shown in Fig. 45. As the critical wavelength is approached the wavelength measured in the guide (λg) may be several times the actual wavelength and hence a small variation in the wavelength to be measured will give a very much larger variation in the wavelength measured in the guide. The guide should be operated near its critical wavelength to obtain accurate measurements. It should be noted that it is important to know the type of mode being propagated in the wave guide, as different modes have different critical wavelengths.

In a practical set-up of this type, the resonance indicator again consists of a crystal detector and microammeter coupled to the guide. The end of the guide is closed by a movable piston driven by a micrometer lead screw. This piston

is adjusted to give several maximum readings on the microammeter. The difference between successive micrometer readings then gives $\lambda g/2$. Knowing the diameter of the guide and the mode being propagated, the actual free-space wavelength may be obtained from the curve (Fig. 45). With a wavemeter of this type, good accuracy can be obtained at wavelengths below 1 cm. (30,000 Mc/s.).

A similar type of wavemeter is the cavity wavemeter, which employs a resonant cavity as the tuning element, variation being obtained by screwing a plunger into the cavity in a fashion similar to the tuning of some types of klystron rhumbatron. Indeed, the tuning of a klystron can be done by using the rhumbatron as a cavity wavemeter. Fig. 46 shows such a wavemeter for the

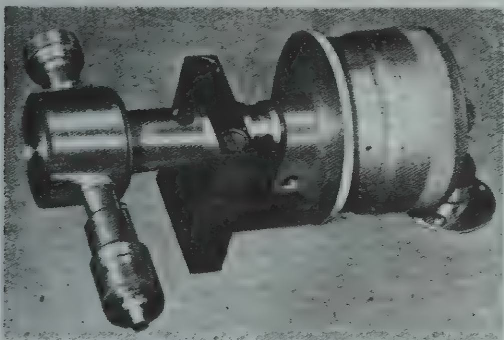


Fig. 46.
Cavity wavemeter for 10,000 Mc/s. *(Photo : courtesy of Marconi Instruments Ltd., St. Albans.)*

3 cms. band. The resonant cavity is at the left of the instrument with the R.F. input socket, connecting to a coupling loop in the cavity, at the rear. The projection on the front of the cavity is the crystal detector, also coupled to the cavity by a loop. A tuning piston is controlled by a fine lead screw, its position being read on the micrometer drum. A calibration has to be used to convert the drum readings to wavelength, an accuracy of about 0·1 per cent. being realised.

Direct frequency measurement can be done by using special types of crystal calibrator. For example, a 300 Mc/s. interpolating oscillator can be set up against a 5 Mc/s. crystal oscillator, and, say, its tenth harmonic used to check a 10 cm. signal. For wavelengths below 10 cms. the experimental technique becomes difficult but instruments are available which go down to 3 cms.

Power

All methods of power measurement at microwave frequencies are direct, *i.e.*, it is power that is measured and not the voltage or current in a known impedance, and they all depend on a thermal effect. For precise, fundamental measurements, the calorimetric method is used and the actual heat evolved when the R.F. is dissipated in a load is the criterion. Less accurate, and incidentally far easier, measurements usually involve measuring the change of resistance of the measuring element when heated by the R.F.

It is well known that a wire filament undergoes a marked resistance change with temperature and special lamps with short straight filaments—bolometer lamps, use this effect. They are made in a variety of sizes from quite large ones for measuring transmitter outputs to tiny ones for measuring milliwatts of power in a wave guide. Arranged so as to absorb all the power to be

measured—in a wave guide by placing it across the guide in the direction of greatest electric field with a backing piston behind—the bolometer lamp is connected in an ordinary Wheatstone bridge circuit. The bridge is balanced with some D.C. flowing through the lamp and the indicating meter brought to the zero position. When R.F. heats the bolometer filament its resistance changes and the bridge unbalance is shown on the meter which can be calibrated in terms of power.

A type of resistance element with a high negative temperature coefficient of resistance is the thermistor which is a bead of particular material, such as uranium oxide, mounted in a suitable envelope. The characteristic is such that at normal room temperature an increase in temperature of about 20° C. halves the value of the resistance. A power of some 60–120 mW. reduces the resistance of a thermistor to 0·2 per cent. of its cold value. Mounted in a wave guide, a thermistor is an invaluable device for monitoring and measuring power. Since they are so sensitive to temperature changes, it is customary to use them in a bridge circuit with other thermistors, external to the wave guide and open to room temperature, as temperature compensators.

Other Measurements

Another important measurement at microwave frequencies is the evaluation of standing wave ratio (S.W.R.). This is the ratio of the power of the forward wave in a transmission line or wave guide and the power of the backwards (reflected) wave and is a measure of the mismatching, etc., caused by obstruction and discontinuities. The measurement is done by using two probes spaced $\lambda/4$ apart in the line and so adjusted that they read respectively the maximum and minimum currents in the line (i max. and i min.). Then

$$s.w.r. = \sqrt{(i\ min./i\ max.)} = (1 - \rho)/(1 + \rho)$$

where ρ is the coefficient of reflection. To secure efficient transfer of power a high S.W.R. is obviously desirable and values not less than, say, 0·95 are sought after. A perfect circuit would give S.W.R. = 1·00.

In receiver work, a variety of measurements is necessary to check performance-sensitivity, noise level and band-width being paramount. Special signal generators are used for the work and are usually equipped with a piston attenuator. This is an inductive or capacitive pick-up element mounted in a section of wave guide such that the operating frequency is well below the critical frequency of the wave guide. Under these conditions as the pick-up is moved away from the exciting source attenuation results and is logarithmically related to the separation so that the attenuator gives a linear calibration in decibels. Modulation of the signal generator is often 50–50 square wave or pulsed with short pulses of about 1 μsec duration.

An undesirable property of pulsed magnetrons is that successive pulses may excite the valve at slightly different frequencies since the magnetron oscillatory circuit has many degrees of freedom. A frequency spectrometer, which is similar to the familiar L.F. visual alignment equipment, allows the complete frequency spectrum of a magnetron to be viewed on a cathode ray tube so that an analysis of its operation can be made. Spectrometers have valuable uses in the laboratory on other microwave applications and their use as a general purpose tool is growing.

These are only a few of the measurements that are associated with microwave apparatus. Many others are routinely made and many others again are desirable. These will not be mentioned as they are already fairly well covered in contemporary literature. Some useful references will be found in the last chapter of this booklet.

CHAPTER 8. BIBLIOGRAPHY

It has been difficult, in view of the amount of space available, to deal at all fully with some aspects of microwave technique and it is suggested that the reader should consult available technical literature for further information. Although the following list of references is by no means complete, it represents a selection covering the main branches of the subject. The references marked with an asterisk can be recommended for general reading.

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INDEX

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FOREWORD

THE purpose of this Publication is to present in as simple a manner as practicable the calculations associated with the application of thermionic valves. It is hoped thereby that the reader will find it possible to utilise the information published by the Valve Manufacturers to the best advantage.

The simplified methods described are sufficiently accurate to obtain optimum results, bearing in mind the usual tolerances of component values and the normal spread of valve characteristics.

It has been assumed that the reader is familiar with the fundamental theory of valves and radio practice as covered in the *Amateur Radio Handbook* and *Radio Handbook Supplement*, and any theory of valves contained herein is in general supplementary.

As far as possible the contents have been confined to valve technique and do not cover in detail any matters properly within the scope of circuit technique. For such information the reader is referred to the other publications in this series.

The co-ordinating authors wish to acknowledge the assistance of Messrs. H. L. Gibson, E. D. Hart, M.A., M.I.R.E., and J. H. Shankland, B.Sc., G8FM, in the preparation of this book.

Glossary of Terms

The following glossary of terms used in this booklet will assist the reader to understand the text without reference to other publications.

A.C.

Alternating Current, of any frequency, but usually applied to power frequencies, *e.g.*, 50 cycles per second.

A.F.

Audio Frequency. Alternating at a frequency within the range of hearing, *e.g.*, between 20 and 12,000 cycles per second.

A.V.C.

Automatic Volume Control. Method employed in a receiver to reduce sensitivity on powerful signals, while leaving the sensitivity at maximum for weak signals.

B.F.O.

Beat Frequency Oscillator. Valve stage employed in a receiver for providing a continuous carrier wave with which the incoming signal beats or heterodynes, producing an audible note, and rendering Morse signals readable.

D.C.

Direct Current. Steady, unfluctuating. Not necessarily a *current*, but merely a steady state, as in " D.C. resistance," " D.C. voltage," etc.

H.F.

High Frequency. Alternating at a frequency between, say, 100 kc/s. and 20,000 kc/s.

H.T.

High Tension. The steady voltage supply which provides the potential applied between anode and cathode of a valve.

I.F.

Intermediate Frequency. The fixed frequency at which most of the amplification in a superhet is carried out.

Q.
Measurement of the " usefulness " of a coil or tuned circuit. The reactance of a coil divided by its resistance at the frequency in use.

R.C.C.
Resistance-Capacity Coupled. Method of coupling a voltage amplifier stage to the succeeding valve. Also referred to as R.C. coupled, or capacity coupled, and sometimes (wrongly) as resistance coupled.

R.F.
Radio Frequency. Alternating at a frequency between, say, 15 kc/s. and 20,000 kc/s. A general term.

R.M.S.
Root Mean Square. The usual method of describing and measuring A.C. voltages. 0.707 times the peak voltage. The R.M.S. value represents the " equivalent " value—*e.g.*, 20 volts A.C., R.M.S., across a resistor would produce the same amount of heat as 20 volts D.C.

V.H.F.
Very High Frequency. (Or U.H.F., Ultra-High Frequency). Alternating at a frequency higher than, say, 30 Mc/s.

Class A
Method of operating a valve so that the grid remains always negative to the cathode. The applied signal voltage is small enough to allow the operating point to remain on the straight portion of the I_a/V_a curve, and no grid-current flows.

Class AB1, Class AB2, Class B
Methods of operating a valve, by progressively increasing both the steady bias voltage on the grid, and the applied signal voltage. The operating point moves beyond the straight portion of the I_a/V_a curve. In Class AB1, no grid-current flows. In Class AB2, grid-current is permitted to flow. In Class B, the steady bias is such that, without any applied signal voltage, the anode current would be reduced to zero. With the signal voltage applied, grid current flows, and the valve is, of course, working beyond the straight portion of the I_a/V_a curve ; because of the steady bias, anode current flows only during the positive half-cycle of the applied signal voltage. Class AB1, AB2 and B all cause distortion of the waveform, and this must be corrected—usually by using two valves in push-pull—but correction is not necessary at radio frequencies, where tuned circuits are used to restore the sine waveform.

Class C
Method of operating a valve with a steady bias more than sufficient to reduce the anode current to zero under no-signal conditions. With the signal voltage applied, anode current flows for only a part of the positive half-cycle. A tuned circuit in the anode converts these " flicks " of anode current into useful power, with current of sine waveform.

Class A, AB1, AB2, B and C
In that order, require progressively greater signal or input-voltages and correspondingly greater input powers, progressively greater steady bias voltages, and, still in that order, they are progressively more efficient—*i.e.*, a valve in Class AB1 provides greater useful output power for the same H.T. supply voltage and current than one in Class A, Class AB2 greater than Class AB1, and so on.

Symbols

The symbols which will be used are, as far as possible, in accordance with those agreed between the *British Radio Valve Manufacturers' Association* and the *British Standards Institution.* Certain other symbols, not included below, are listed in the particular section in which they are used.

Voltage

V_a	D.C. anode voltage.
$V_{a(b)}$	D.C. H.T. supply voltage.
V_g	D.C. grid voltage.
V_{g_1}, V_{g_2}, V_{g_3}, *etc.*	D.C. voltages applied to grids numbered outwards from the cathode.
V_{sig}	Signal voltage.
V_{out}	Output voltage.

Current

I_a	D.C. anode current.
I_g	D.C. grid current.
I_{g_1}, I_{g_2}, *etc.*	Currents to grids Nos. 1, 2, etc.

Resistance

R_a	External anode load resistance.
R_g	External grid leak.
R_{g_2}	External resistance in series with g_2.
R_k	External cathode resistance.

Impedance

r_a	Internal anode impedance.
Z_a	External load impedance.

Capacity

c_{in}	Input capacity (grid to all electrodes less anode).
c_{out}	Output capacity (anode to all electrodes less grid).
c_{ga}	Capacity of grid to anode.

Miscellaneous

μ	=	Amplification factor.
ω	=	$2\pi \times$ frequency.
L	=	Inductance.
W_{out}	=	Power output.
w_a	=	Anode dissipation.
g_m	=	Mutual conductance.
g_c	=	Conversion conductance.
f	=	Frequency.
Δf	=	Total circuit band width.

CHAPTER 1 GENERAL CONSIDERATIONS GOVERNING VALVE OPERATION

IT is proposed in this chapter to give some guidance regarding the considerations which govern the operating conditions of thermionic valves.

It is not generally understood why valve manufacturers are so particular about the conditions of use. The published ratings are believed by many to be unnecessarily conservative. It is hoped that a study of the subject, as dealt with in the following pages may result in clearing up some of these doubts.

Types of Cathodes

Originally all valves used pure tungsten—operating at very high temperatures, as an electron source—but since the discovery of the improved performance obtained by the introduction of certain additional agents the use of this metal has been discontinued for all receiving and small transmitting valves.

If, in a given valve, the emission current is determined by its geometry and electrode potentials, rather than by the cathode emission available, the current is said to be " space-charge limited " and the emission current will be independent of slight changes of cathode temperature. If the electrode potentials are adjusted so as to tend to draw-off more current than the cathode can supply, the current is said to be " temperature limited " ; any slight change in cathode temperature results in an immediate change in emission current.

Pure Tungsten

Pure tungsten has a definite maximum emission, dependent upon its temperature, and owing to the non-employment of any activating addition agent cannot be " poisoned " or de-activated. It may, therefore, be run under temperature-limited conditions without ill effect and it will not be damaged by over-heating, providing the period of over-heating is not long enough to produce a material reduction in the cross-section of the filament, by simple evaporation of the metal.

Pure tungsten operates at a temperature of about 2300°C. with an emission of about 4 milliamperes per watt of cathode heating.

Thoriated Tungsten

By the correct addition of thoria and carbon to tungsten, an emitting surface of greatly increased efficiency results. This is known as " thoriated tungsten." Thoriated tungsten runs at a temperature of approximately 1650°C. and has an emission efficiency of about 30 milliamperes per watt. The main factor in achieving this improvement is a very thin layer of thorium on the cathode surface.

During valve manufacture some of the thoria will have been reduced to thorium within the filament core, the thorium then diffusing outward to form a surface film. Under running conditions there is a slow loss of thorium due to evaporation, but the rate of diffusion should keep pace with it, maintaining a constant surface. The rate of loss of thorium is somewhat accelerated by drawing an emission current. A poor valve may be unable to cope with this

condition and the emission will " sag " when H.T. is applied, recovering again during " filament only " stand-by periods. This sagging performance is typical of " below par " thoriated tungsten valves. Some improvement may be achieved by a slight increase in filament volts; the rate of diffusion may increase more rapidly than the loss, and, due to the higher temperature, the emitting properties of the layer will as a consequence be improved. A large increase in temperature however will increase the loss-rate, and a smaller fraction of the surface will be coated, whilst above 2200°C. the cathode will behave like one of pure tungsten. Any increase in temperature will result in some shortening of valve life for, although the relative diffusion rate may be increased the absolute loss-rate will also be increased.

It is sometimes possible to re-activate a valve by " flashing " at about 2·5 times normal voltage for 1 minute. This operation, which reduces more thoria to thorium, should be followed with a 15 to 30 minutes run at about 1·5 times normal filament voltage, to give the new thorium a chance to diffuse to the surface. Should there be no thoria left to reduce, this treatment will " flash-off " the existing thorium with no means of replacing it and the valve will be useless.

Thoriated tungsten valves should not be run continuously in a temperature-

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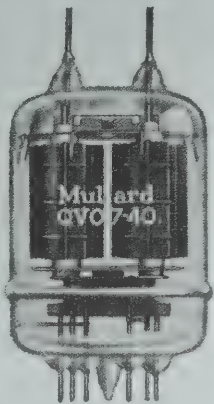


limited condition, as this tends to be destructive to the cathode surface, but an occasional excursion into this region may not be harmful. For instance, in an anode-modulated Class C stage, temperature limitation can occur at peak-positive grid-voltage and peak-positive anode-voltage for 100 per cent. modulation without harm. This is possible mainly, because 100 per cent. modulation occurs only occasionally in practice. Although no harm to the valve results there will be some distortion and decrement modulation. The appearance of decrement modulation in an anode modulated stage that has previously been satisfactory is often due to incipient low emission of the power amplifier. If manufacturers' data on safe emission is not available the following information should enable a safe design figure to be obtained.

Average D.C. cathode current maximum, 7·5 mA per watt of cathode heating.

Peak cathode current maximum, 30 mA per watt of cathode heating.

Oxide-Coated Cathodes

The oxide-coated cathode represents an advance on thoriated tungsten even greater than the latter shows over pure tungsten. The operating temperature has been reduced to approximately 720°C., and the emission figure raised to 200 milliamperes per watt. The cathode coating is formed from a paste of the oxides of the alkaline earth metals, mostly barium, applied by dipping or spraying. The process is equally suitable for directly or indirectly-heated cathodes whereas the previous types were confined to the directly-heated case.

The emission is regarded as arising from a film of metallic barium on the surface of the oxide layer. The free barium is formed by reduction of the oxide and is brought to the surface by diffusion. Thus the mechanism is similar to that of thoriated tungsten and there are similarities in the activation treatment. Barium is very active chemically and the thin film on the cathode surface can easily be poisoned by traces of gas liberated internally by incorrect operation. Valves that have failed prematurely, due to surface poisoning, can sometimes be repaired by flashing, *i.e.*, evaporating the damaged layer and re-forming some more barium, and then running to diffuse the barium to the surface. If a valve has had a reasonable life and is failing due to loss of barium rather than poisoning this treatment will accelerate the decline. The suggested treatment is as below :

(*i*) Flash for 1 minute at 2¼ times normal filament voltage.

(*ii*) Run for 30 minutes at 1¼ times rated filament voltage with rated voltages applied to all other electrodes.

(*iii*) Run for 1 hour at rated filament volts with no other potentials applied.

Although the oxide-coated cathodes should never be run under temperature-limited conditions, no harm results from running at considerably reduced temperatures providing the emission demand is also reduced. For example a resistance-capacity-coupled pentode audio-voltage-amplifier may have its total cathode current reduced to perhaps 0·5 milliampere by the high value of anode and screen resistors normally used. Under these conditions most valves will work down to 70 per cent. of normal filament voltage giving prolonged life and a reduction in hum and thermal noise. If the filament voltage is raised above normal the emission, and particularly the peak emission, will be increased, but the life will be shortened due to the greater rate of barium evaporation. Further, since some of the barium will be deposited upon the grid, there is an increased chance of developing grid-emission which is emphasised by the increased grid temperature resulting directly from the increased filament wattage.

Some valves intended specifically for Class C or other operation demanding high peak emission, are designed with abnormally hot cathodes, the shorter life expectancy being off-set by the improved performance. The peak emission that can be drawn depends upon the duration of, and space between, the peak demands and where manufacturers' figures are not available the following information can be taken as a guide.

Average D.C. cathode current maximum 20 mA per watt.

Class B operation, conductivity nearly 50 per cent. of time, peak current maximum 50 mA per watt.

Class C operation, conductivity 10 per cent. to 20 per cent. of time, peak current maximum 200 mA per watt.

Pulse operation, conductivity 0·1 per cent. to 0·5 per cent. of time, peak maximum 500 mA per watt.

Electrode Dissipation

Most metals, and many other substances used in valve making, have the power of absorbing large quantities of various gases and it is the manufacturers' concern to ensure that all parts are treated so as to prevent evolution of this gas during life; the hotter the part runs, the more vigorous and lengthy the treatment has to be. From the cost point of view it is often better to add radiating fins in order to reduce the running temperature of an electrode rather than to pay for more complicated treatment processes. To safeguard his own reputation, a manufacturer will treat his components so that there is a fair margin between the maximum rating and the point of serious gas release, but economic considerations dictate that this margin shall not be large. Any over-dissipation by an electrode will, therefore, entail a risk of gas evolution which may either poison the emitting surface or partly destroy the vacuum.

Manufacturers' data is usually very definite in specifying maximum dissipation and it should not be exceeded. In the rare cases where no data is available, any electrode, other than a filament or cathode, that shows a visible glow must be regarded as overloaded, and dissipation must be reduced. Among valves of the transmitting type there are some using tantalum components. Tantalum, unlike most other metals tends to absorb gas rather than evolve it when operating at high temperatures. Valves using this material are usually designed so that for normal dissipation the anodes run at a bright orange temperature. These are obvious exceptions to the " no glow " rule.

Negative Grid Current

It is well-known that the flow of grid current through a grid resistor produces a negative bias on the grid and it follows that if the direction of the grid current is reversed, the effect of the potential drop across the grid resistor would be to produce a positive bias, or at least reduce any negative bias already present, which will in turn result in increased space current and dissipation.

Such a reversed or negative grid current can arise from three causes:

(*a*) Internal leakage paths within the valve or its base.

(*b*) Positive ion current, due to faulty vacuum.

(*c*) Electron emission from the grid (grid emission).

All these factors tend to increase with increase in temperature, and as increased temperature will result from any increase in dissipation the defect, once it has developed, tends to maintain and increase itself. However, where cathode bias is used, any increase in cathode current provides automatically an increase in bias so that in this case there is a slight compensating action.

The existence of this type of trouble will show itself as excessive anode current (leading to damage of the valve or the power supply) and/or heavy damping of the input circuit, such that the previous stage will appear to have lost most of its amplification.

The damage can be avoided by attention to the following points:

(*i*) Do not use excessive filament voltage. Never exceed 106 per cent. of the rated value.

(*ii*) Do not permit excessive dissipation.

(*iii*) Limit the total D.C. resistance of the grid-cathode path to a safe value.

The first two precautions should be obvious, whilst condition (*iii*) can be

met by studying the manufacturers' data. When no data is available, the following figures are suggested as a guide:

For fixed bias.

Total cathode current (in milliamperes) $\times$ grid resistance (in megohms) < 15.

For cathode automatic bias.

Total cathode current (in milliamperes) $\times$ grid resistance (in megohms) < 25.

By the above rules where a resistance is included in the grid path of two or more valves (such as the decoupling and diode load resistors of A.V.C. systems), its value must be multiplied by the number of common paths in order to ascertain the equivalent resistance of the grid path in any one valve.

Heater-to-cathode potential

It is good practice to keep the potential difference between the heater and cathode of an indirectly-heated valve as low as possible. Where the maximum rating is specified it should be observed; in other cases the potential in ordinary types of valves should not exceed 50 volts. Modern valves have a very high order of heater-to-cathode insulation. For example, rectifiers in series-run apparatus may have an average of 250 volts (with peaks of over 500 volts) between heater and cathode, and great liberties may sometimes be taken with ordinary valves without deleterious results.

High heater-to-cathode potentials will not cause any slow deterioration of the valve. Either the insulation will stand the voltage or breakdown will occur leaving a permanent short-circuit between heater and cathode or a broken filament. Breakdown is more likely to occur during warming up, and its possibility can be reduced if the application of heater-to-cathode potential is delayed until the cathode has attained operating temperature.

Some leakage current between heater and cathode may flow even with quite low applied voltages and increasing this voltage will not increase the current necessarily in proportion. The leakage current will also be different in value depending on the direction of the applied voltage. In certain types of valve it is possible for the heater to emit to the grid, causing hum, and if this occurs the hum can be reduced by raising the heater-cathode voltage so that the heater is positive with respect to the grid and cathode; generally a D.C. voltage of the order of $+5$ to $+ 10$ volts applied to the centre tap of the heater supply is all that is necessary.

If it is necessary to open the cathode circuit of a valve for the purposes of keying or muting, a resistance not exceeding 0·25 megohms should be permanently wired between cathode and heater to avoid subjecting the insulation to the full anode or screen potential.

The heater-to-cathode insulation if included across any tuned circuit may give rise to modulation hum and frequency instability because the insulation will vary with temperature and hence with heater voltage, and the capacity will alter with movement of the heater within the cathode.

Switching Sequence

The application of heater and anode voltages simultaneously results in a valve operating for a short period under temperature-limited conditions during the warming-up process. For this reason it is advisable to delay the application of H.T. voltage until the cathodes have attained normal temperature. This complication is not worthwhile for vacuum valves operating at less than 400 volts. Delayed switching is desirable for higher voltages and *essential* for gas-filled or mercury-filled valves.

CHAPTER 2 VOLTAGE AMPLIFIERS

THE following additional symbols are used in this Chapter.

R_o .. .. .. Succeeding valve grid leak.
C_k .. .. .. Cathode by-pass condenser.
C_o .. .. .. Inter-stage coupling condenser.
C_{g2} .. .. .. Screen by-pass condenser.

Voltage Amplifiers

When it is desired to use a valve primarily for the purpose of increasing the signal voltage, such as in an audio or I.F. amplifier, as distinct from increasing the power, such as in an A.F. or R.F. output stage, there are four possible methods, *i.e.* resistance-capacity coupling, choke-capacity coupling, tuned anode or tuned grid coupling and transformer coupling.

In general, the first is suitable only for audio frequencies unless the gain per stage is very low and the output voltage required fairly small. The second is suitable for somewhat higher frequencies, providing the choke is suitably chosen. The last two may be used up to very high frequencies, the limitation being the tuned circuit, the transformer and the valve itself.

Somewhat different considerations apply to these methods when pentodes are used instead of triodes. The necessary procedure to employ, in order to design such amplifiers, is described below.

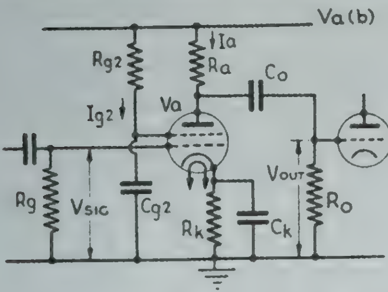


Fig. 1.
Typical basic circuit of a resistance capacity coupled voltage amplifier employing a tetrode or pentode valve.

Resistance-Capacity Coupled (R.C.C.) Amplifiers

The conventional circuit of an amplifier is shown in Fig. 1. R_g is the grid leak across which the input voltage V_{sig} is applied. R_a is the anode load resistor to which the H.T. line voltage $V_{a\ (b)}$ is applied. V_a is the anode voltage on the actual anode. R_k is the cathode bias resistor by-passed by the decoupling condenser C_k. These are omitted in the case of directly heated battery valves. C_o is the coupling condenser and R_o is the grid leak of the succeeding stage across which the amplified output voltage V_{out} is required.

If the amplifier valve is a pentode or tetrode and not a triode then there is a screen dropping resistor R_{g_2} by-passed by the condenser C_{g_2}, the actual screen voltage being V_{g_2}.

The first essential is to choose the type of valve. The considerations are: – the output voltage required, the H.T. voltage available, the value of the succeeding valve grid leak and the highest frequency to be amplified. If, for example, the succeeding valve is an output stage it will require an input signal between say 5 volts and 50 volts or more, depending on the type, whereas if it is a further voltage amplifier, the input is not likely to exceed 10 volts and almost any valve will furnish this voltage. The H.T. line voltage will usually be fixed by the design of the remainder of the equipment, but as a general guide this voltage must not be less than about four times the output voltage V_{out} required, and preferably not less than six times. As long as the actual anode voltage V_a does not exceed the makers' rating all is in order.

The value of the succeeding stage grid leak R_o is mainly a question of the type of valve used. Output stages should not as a rule have grid leaks higher than 100,000 ohms unless they are used with automatic grid bias in which case up to 500,000 ohms may be used, whereas small valves (as used in voltage amplifiers) may use a megohm or higher. The coupling condenser C_o should be so chosen that it has a low reactance compared with the value of R_o at the

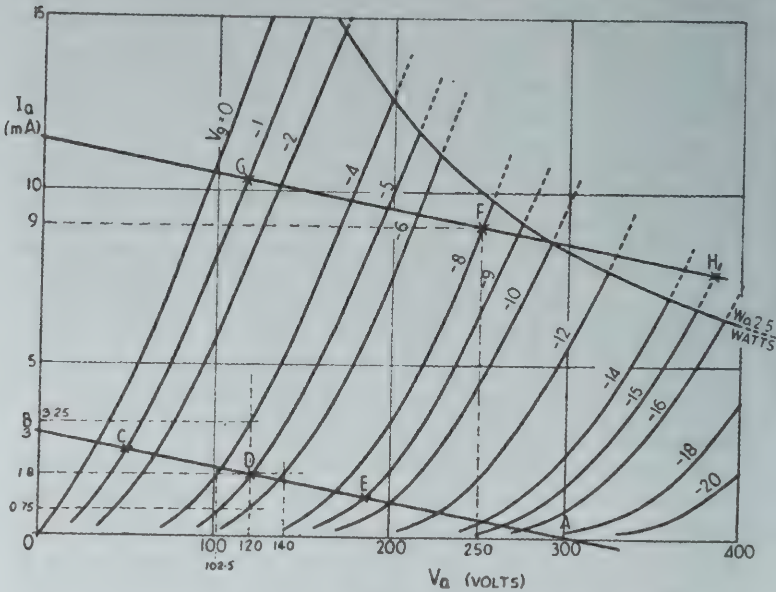


Fig. 2.

Typical characteristic curves I_a/V_a of a triode valve showing load lines when used as a voltage amplifier.

lowest frequency to be used, a suitable ratio being from one half to one quarter. The highest frequency to be used is relevant, because across R_a or R_o are the capacities of the anode of the amplifier valve, the input capacity of the succeeding valve and the wiring capacities. The sum of these capacities being in parallel with R_a (which is itself in parallel with the anode impedance of the valve r_a) will lower the effective anode load impedance and hence the gain of the stage. For audio frequencies with good high frequency response the value of R_a should not exceed 1 megohm and preferably be less, since the total capacity across it will vary from about 30 pF to possibly 100 pF, giving a shunt reactance of 500,000 ohms or less. The effect of the shunt capacities is more important if a pentode is used as the voltage amplifier than with a triode because the pentode anode impedance is much higher.

If the succeeding valve is a tetrode or pentode its input capacity will be less due to reduced "Miller" effect. If a pentode is employed the screen by-pass condenser C_{g_2} should have a value such that its reactance is low compared with the apparent screen impedance. This impedance is approximately the screen voltage V_{g_2} divided by the screen current I_{g_2}.

Triodes

In order to ascertain the values to be used with a triode valve the essential characteristic curve is that obtained by plotting anode current against anode voltage (I_a/V_a) a typical example of which is shown in Fig. 2.

Let it be assumed that the H.T. line voltage ($V_{a(b)}$) is 300 volts, the anode load resistor is 100,000 ohms and the succeeding valve grid leak 500,000 ohms. It is required to find a suitable value for the cathode resistor, the voltage amplification and maximum available output without distortion.

Since with a grid bias of more than cut-off value, the anode current will be zero, the voltage drop in the anode load R_a will be zero. Hence one end of the load line will be represented by 300 volts at point A and the slope of the line will be 100,000 ohms. Hence the other end will be at 3 mA at point B, and as we vary the grid voltage the anode voltage V_a will travel along this line A B. In general, the operating point will be such that the mean anode voltage V_a is between 0·4 and 0·5 times the H.T. line voltage. The anode voltage can thus swing above and below this value. The figure 0·4 should be used for triodes having an amplification factor lower than 40 and 0·5 for those having an amplification factor higher than 40 since the latter have a shorter grid base, and in consequence the operating point is closer to the point at which grid current starts. In the average type of valve grid current starts at just under 1 volt and therefore the grid voltage should never swing to the left beyond –1 volts (the point C on the curve). As these curves are those of a triode with a nominal amplification factor (μ) of 20, then 0·4 times 300 volts = 120 volts (point D on the curve). This corresponds to a grid bias of –5 volts and an anode current of 1·8 mA. Hence the cathode bias resistor value is 5,000/1·8 = 2,800 ohms. In practice 3,000 ohms could be used.

If we follow the vertical line up and down through point D we see that the anode current is 3·25 mA at –4 volts and 0·75 mA at –6 volts, hence for 2 volts change in grid voltage there is 2·5 mA change in anode current so that the mutual conductance at this point is 1·25 mA/V. If we follow a horizontal

line through point D either way we see that –4 volts corresponds to 102·5 volts and –6 volts corresponds to 140 volts. Thus 2 volts change in grid voltage is equal to 37·5 volts change in anode voltage. Hence the amplification factor (μ) at the operating point is 18·75.

From this it follows that the anode impedance is

$$\frac{18 \cdot 75}{1 \cdot 25} \times 1{,}000 = 15{,}000 \text{ ohms.}$$

The actual output load at audio frequencies is R_a in parallel with R_o which is 100,000 in parallel with 500,000 or 83,000 ohms.

The voltage gain is $\dfrac{\text{amplification factor } (\mu) \times \text{output load}}{\text{output load} + \text{anode impedance}}$ which is

$$\frac{18 \cdot 75 \times 83{,}000}{83{,}000 + 15{,}000} = 16.$$

Hence for every 1 volt R.M.S. applied to the grid we shall have 16 volts available for the succeeding valve.

Since the grid can only swing to point C as mentioned above, the peak input voltage is 4 volts so that the maximum output V_{out} is 64 volts peak or 45 volts R.M.S. The values of the condensers are chosen as mentioned in a preceding paragraph.

In order to calculate the distortion the voltage gain must be re-calculated in just the same way about the point C and also at the point E (–9 volts) which is at the equivalent distance on the other side of the bias voltage. These work out to be 16·3 at C and 14·3 at E, hence the positive peaks of the signal output will be amplified less than the negative peaks. Naturally if a smaller input swing is employed the distortion will be correspondingly less.

The input capacity of the stage will be the grid-cathode capacity plus the product of the grid-anode capacity and the voltage gain, which in the above case is very little less than the amplification factor.

Pentodes or Tetrodes

The choice of a pentode or tetrode, instead of a triode, will depend on the gain required and upon the load presented by the succeeding valve. Valves of these types, due to their much higher amplification factor, will give considerably more stage gain but in general will handle a smaller signal input without distortion. Their higher output impedance makes them unsuitable unless negative feed-back is applied, for driving any output stage that is operated other than strictly Class A1. These factors indicate that they are most suitable for early stages in a multistage amplifier, but in this connection it is as well to bear in mind that they also are inherently more noisy than triodes.

In order to determine the operating conditions the treatment has to be different from that of a triode because of the complications due to the valve performing as a constant current device and the additional factor of screen current.

As an example the anode current, anode voltage (I_a/V_a) curves for a typical pentode are shown in Fig. 3. The rated Class A amplifier conditions are V_a250, V_{g_2} 100, V_{g_1} –3. Using an anode load of say 500,000 ohms, the load line will be AB. It is evident that the anode voltage at the operating point C will be very low and further, the anode will be unable to swing downwards in a linear manner, although it will swing upwards.

These conditions, whilst desirable for a limiter are not those for a linear amplifier. The trouble in this example is that the screen voltage is too high. Fig. 4 shows I_a/V_a curves for the same valve with a screen voltage of 30 volts. The load line AB for 500,000 ohms gives the operating point of C at an anode voltage of 90 which is just under half the H.T. line voltage, with a grid voltage of $-1{\cdot}5$ volts. Under these conditions the valve would operate satisfactorily as an R.C.C. amplifier. It would appear that all that is required is a curve at the correct screen voltage, but the difficulty is that valve manufacturers do not supply a folder of curves at all values of low screen voltage.

The data required is a set of dynamic curves on which is plotted anode current against grid voltage for various screen voltages with given values of anode load. Such a curve for the above valve is shown in Fig. 5. This curve is plotted with an H.T. line voltage ($V_{a\ (b)}$) of 250 volts and an anode load (R_a) of 500,000 ohms. It will be noticed that for every screen voltage a curve is obtained, which has a straight portion in the centre, but is bent at either end, the slope of the curves getting progressively less as the screen voltage is increased. For every screen voltage there is a correct grid voltage, but for the best stage gain a grid voltage should be chosen such that the working part of the curve does not extend to the right beyond –1 volt because of grid current. From the above, the correct curve is that corresponding to a screen voltage of 30 volts. The operating point is approximately in the centre of the straight portion of this curve, corresponding to point A, the grid voltage

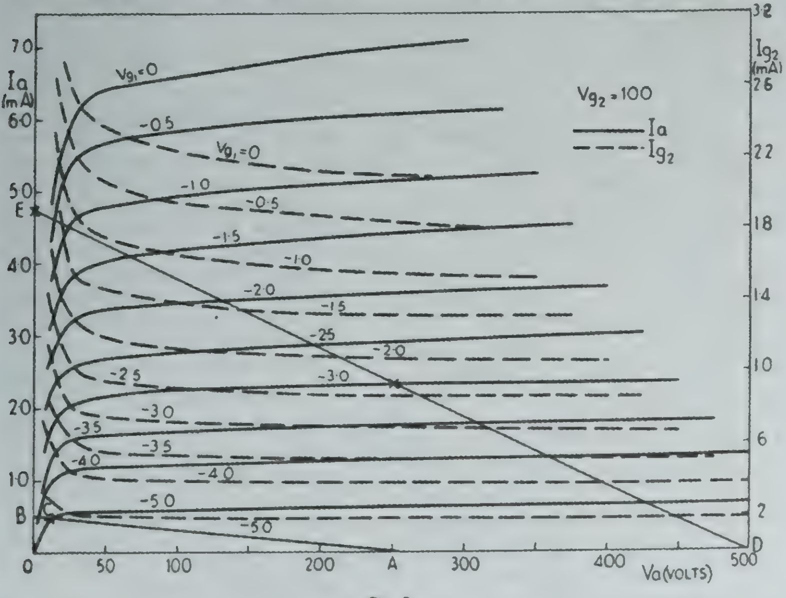


Fig. 3.

Typical characteristic curves I_a/V_a of a R.F. Pentode valve showing load lines when used as a voltage amplifier.

being $-1{\cdot}5$ volts, and the anode current $0{\cdot}31$ mA. Hence the anode voltage (V_a) is 94 volts. The dotted curves indicate the screen current, which at screen volts 30 and grid volts $-1{\cdot}5$ is $0{\cdot}085$ mA. Hence the bias resistor (R_k) is

$$\frac{\text{grid volts}}{\text{anode + screen current}} \text{ or } \frac{1{\cdot}5 \times 1{,}000}{0{\cdot}31 + 0{\cdot}085} = 3{,}800 \text{ ohms}$$

and the screen resistor (R_{g_2}) is

$$\frac{\text{H.T. line voltage} - \text{screen voltage}}{\text{screen current}} \text{ or } \frac{250 - 30}{0{\cdot}085} \times 1{,}000 = 2{\cdot}6 \text{ megohms.}$$

Since the anode current is flowing through the load resistance there is a linear relation between the anode current and anode voltage. If points are taken either side of the working point such as grid volts $-1{\cdot}75$ and $-1{\cdot}25$ volts (points B and C) these correspond to anode currents of $0{\cdot}2$ mA and $0{\cdot}43$ mA, and (since these currents flow through the 500,000 ohms load) to anode voltages of 150 and 35 volts. Therefore a change of $0{\cdot}5$ volts on the grid produces 115 volts change in the anode voltage, hence the gain is 230 times. This figure of course, presumes that the cathode and screen are adequately decoupled as mentioned earlier. The net gain is less than this because, as in the triode case, the succeeding valve grid leak R_o has to be taken into account. If the grid leak R_o is taken as 1 megohm in this case, the gain will be reduced by $\frac{R_o}{R_o + R_a}$ or $\frac{1}{1 + 0{\cdot}5}$ or $0{\cdot}67$, *i.e.* the net gain is 154 times.

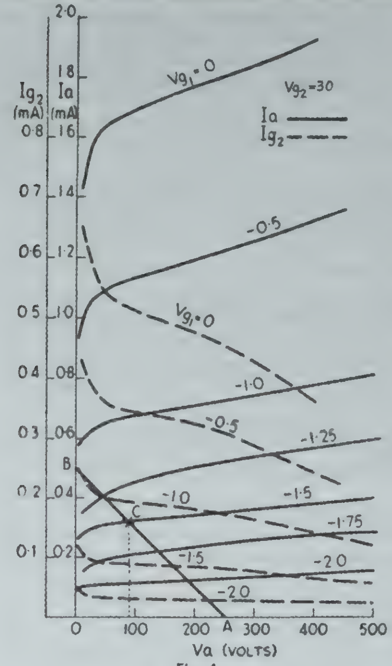


Fig. 4.

I_a/V_a characteristic curves for the R.F. Pentode shown in Fig. 3 but with screen voltage V_{g_2} = 30 volts.

The maximum output we can achieve without excessive distortion is the linear part of the curve E F, which corresponds to a peak to peak anode current change of $0{\cdot}35$ mA, or 175 peak to peak voltage change or 62 volts R.M.S.

As a fairly reliable guide the maximum output voltage is obtained when the anode current is so chosen that it equals

$$\frac{0{\cdot}56 \times \text{H.T. line voltage } (V_{a\,(b)})}{\text{anode load resistor } (R_a)}$$

The distortion can be obtained in the same way as for triodes by working out the stage gain from the curves at the extremes of the grid swing, the anode impedance being neglected, as in general it is high compared with the likely values of anode load.

When such curves cannot be obtained from the valve manufacturer it is not difficult to set up a

suitable circuit and take curves of the anode current/grid voltage for a few selected screen voltages and from this data the operating point can be selected. The screen current may be measured at the operating point only, in order to determine the screen dropping resistor and the bias resistor.

Choke-Capacity Coupled Amplifiers

It is evident that if a choke is substituted for the anode load (R_a) in Fig. 1, the circuit will operate in the same manner, but since there will be little D.C. voltage drop in any well designed choke, the anode voltage will be equal to the H.T. line voltage ($V_{a(b)}$). Normally the impedance of the choke should be very high compared with the valve impedance (r_a) and the succeeding valve grid leak (R_o) over the whole frequency range at which it is desired to amplify. If this is not so then the load for the valve is reactive and the load line takes the form of an ellipse, which becomes a circle when the resistance of the grid leak is infinitely high compared with the reactance of the choke. The centre of the circle or ellipse is the normal operating point. This effect is caused by the self inductance of the choke causing the anode voltage to be out of phase with the current through the choke.

Normally, the above remarks only apply at the lower extremes of the frequency range where the reactance of the choke is falling rapidly or in certain special applications, such as television line time-base amplifiers, where the load due to the deflector coils is almost wholly reactive.

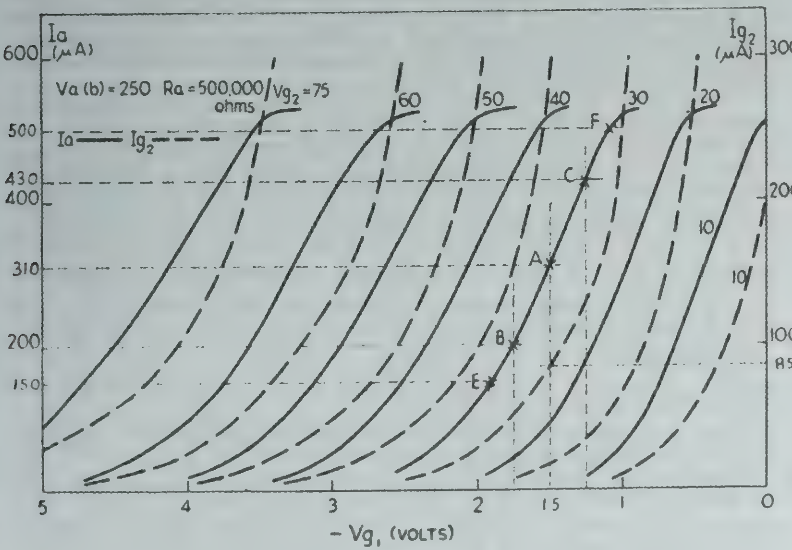


Fig. 5.

I_a V_{g1} curves for the R.F. Pentode shown in Fig. 3 but plotted with an anode load R_a of 500,000 ohms and varying screen voltage V_{g2}.

Taking a practical case when the choke has an adequately high reactance and can be ignored, the load is then the grid leak of the succeeding stage, say 100,000 ohms. Then if the triode valve shown in Fig. 2 is used and an H.T. line voltage of 250 volts is employed, the anode voltage will be almost 250 volts if the D.C. drop in the choke is small. If the makers' rating for a Class A amplifier at this voltage is grid volts —8, then the operating point will be F and the load line GH is drawn with a slope of 100,000 ohms. This line extends to the right to a potential nearly equal to twice the H.T. line voltage, the source of this voltage being the magnetic field in the choke, which, when the anode current is falling, resists the fall in flux and in doing so induces in the winding a voltage which is additive to the supply voltage.

Since the grid bias is –8 volts, the maximum input swing is from –1 volts (G) to –15 volts (H). The mutual conductance at F determined in the manner described before, is 2·5 mA/V and the amplification factor 21, hence the anode impedance (r_a) is 8,300 ohms. The voltage gain is therefore

$$\frac{21 \times 100{,}000}{8{,}300 + 100{,}000} = 19{\cdot}4.$$

and since the voltage input swing is 14 volts peak-peak (or 5 volts R.M.S.) the output will be 97 volts R.M.S.

It is seen that the gain is higher than in the resistance-capacity coupled case and the output voltage is also increased. The cathode bias resistor is obviously 1,000 × 8/9 ohms or 900 ohms.

If, for example, the stage is used at audio frequencies and the choke is iron cored, its inductance will vary with frequency. Knowing this relation the gain at different frequencies we can calculate, and find the drop in gain at the low frequency end. The effective load will be the reactance of the choke in parallel with the grid leak of 100,000 ohms used in the example. It must be borne in mind that the reactance of the choke must be added *vectorially* not directly to the resistance.

When a pentode is used instead of a triode as a choke-capacity coupled stage, the procedure is the same as that for a triode, because we can now use the normal anode current/anode voltage curves, the anode voltage being equal to the H.T. line voltage. As an example, a load line DE has been drawn on the curves in Fig. 3 equivalent to a load of 100,000 ohms.

Tuned anode coupling, such as is used in the R. F. stages of a receiver, is a variation of a choke-capacity coupling. In this case, the choke is replaced by a tuned circuit whose dynamic resistance $\frac{\omega^2 L^2}{r}$ (or Reactance × " Q ") may be treated as being in parallel with the succeeding valve grid leak, and the calculation becomes similar to that for a resistance-capacity coupled amplifier.

Transformer Coupled Amplifiers

When a transformer is used for coupling between stages the procedure is the same as for choke coupling. The transformer will merely step up the voltage from the primary into the secondary, thus increasing the stage gain by the radio of the transformer turns. For example, a 4 to 1 step up transformer used instead of a choke in Fig. 2 where the stage gain was shown to be 19·4, would increase the gain by four times to approximately 80 and the output voltage correspondingly. The output load is then the primary impedance of the transformer in parallel with the reflected secondary load. The reflected secondary load is the load on the secondary divided or multiplied

by the square of the primary to secondary turns-ratio depending upon whether the transformer is a step up or down ratio. Generally there is no resistive secondary load, but there may be considerable capacity load at the higher frequencies or there may be a grid current load in the case of a Class B or AB2 amplifier. The transformer used may be a tuned transformer such as an I.F. transformer, and similar procedure is used if the dynamic resistance of the transformer is known and used as the load for drawing the load line on the I_a/V_a curves. If the transformer is of high dynamic resistance allowance must be made for the effect of valve anode impedance on both the gain and the selectivity.

Wide-Band Amplifiers

For a number of purposes a voltage amplifier is required having a very wide band-width. Examples of such use are cathode ray oscilloscopes and television. These amplifiers may be of any of the types mentioned above, but generally they will use pentodes because of the increased gain and the greater stability afforded by the lower grid-anode capacitance.

Wide-Band Video Amplifiers

It will be remembered that frequency limitations are imposed on a resistance-coupled amplifier by the sizes of the coupling condensers at the low frequencies, and the capacity across the output load at the high frequencies. The low frequency gain may be maintained by the use of larger coupling condensers, but a limit is reached when these become unwieldy and it becomes difficult adequately to decouple the H.T. supply and cathode resistor at these low frequencies. If frequencies down to D.C. are required there is no alternative but direct coupling. It is not proposed here to deal with direct-coupled amplifiers because the treatment as regards the drawing of load lines and calculations of gain is not affected in any way by the method of coupling.

The high frequencies in a normal resistance coupled amplifier are attenuated by the capacities, so that using conventional resistor values the upper frequency limit is in the region of 25 kc/s. or lower. But if the values of the anode load (R_a) or succeeding valve grid leak (R_g) are lowered, then the shunt capacity becomes less important ; at the same time, of course, the voltage gain and output voltage are reduced.

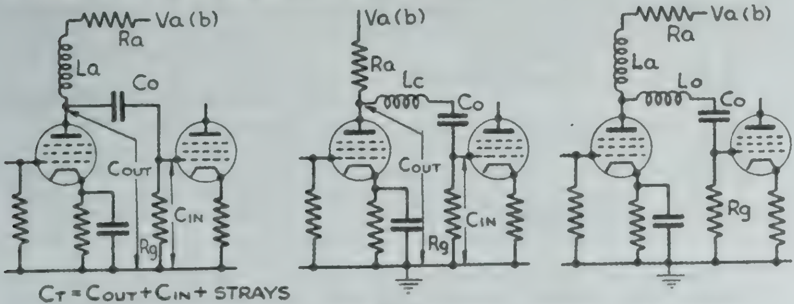


Fig. 6.
Three methods of overcoming the drop in high frequency response of wide band vidio amplifiers.
(a) Shunt compensation. (b) Series compensation. (c) Shunt series compensation.

For this reason it is usual to employ pentodes having a very high mutual conductance of the order of 7·5–10 mA/V and even then the stage gain is not very great.

When a low anode load (R_a) is employed the voltage gain becomes nearly equal to the mutual conductance multiplied by the load, because the valve impedance is infinite by comparison. For example if R_a is 5,000 ohms, r_a is 0·3 megohm and g_m is 10 mA/V then the amplification factor μ is 3,000.

Voltage gain $= \dfrac{3{,}000 \times 5{,}000}{5{,}000 + 300{,}000}$ which is nearly $\dfrac{3{,}000 \times 5{,}000}{300{,}000}$ or 50

and as $\dfrac{\mu}{r_a} = g_m$, the expression $= g_m \times R_a$ applies. Obviously, since the capacities across the load are the output capacity of the valve, the input capacity of the succeeding valve and the circuit strays, it is desirable for these to be as low as possible and the best valves to use for wide band-width are those having the highest ratio of g_m to capacity. The upper frequency limit (f_{max}) with R.C. coupling is usually defined as the point at which the response is 0·707 of the value of the level response at lower frequencies and this occurs when $R_a = \dfrac{1}{2\pi f_{max} C_t}$ where C_t is the total capacity across the load in Farads.

For example, using a R.F. pentode having a slope of 7·5 mA/V and an input capacity c_{in} of 7·5 pF an output capacity c_{out} of 3·5 pF and assuming circuit strays of 5pF with an anode load of 3,300 ohms, the voltage gain will be $\dfrac{7{\cdot}5 \times 3{,}300}{1{,}000} = 25$, and the maximum frequency

$$f_{max} = \frac{10^{12}}{2\pi \times 3{,}300 \times 16} = 3 \text{ Mc/s.}$$

If the valve capacities or the strays were higher (for example a total of 25pF) the maximum frequency would be reduced to 1·9 Mc/s.

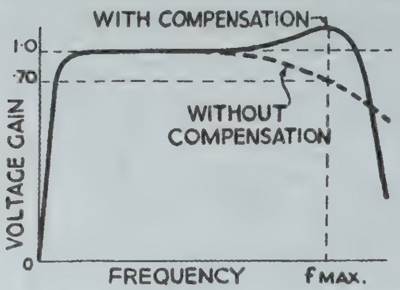


Fig. 7.
Curves of the frequency response of a wide-band vidio amplifier with and without compensation.

The effect of the capacities may be reduced by using a small amount of inductance in the load circuit. The three methods of achieving this are known as "Shunt compensation (Fig. 6a), "Series compensation" (Fig. 6b), and "Shunt-series compensation" (Fig. 6c). The curves shown in Fig. 7

indicate the response with and without compensation. The design figures for each are as follows :—

" Shunt "

$$R_a = \frac{1}{2\pi \times f_{max} \times C_t}$$

where R_a is in ohms f_{max} is in cycles C_t is in farads.

$L_a = 0{\cdot}5 \times C_t \times R_a^2$ where L_a is in henrys.

Relative gain at $f_{max} = 1{\cdot}0$.

" Series "

($2 \times c_{out}$ should be adjusted to equal c_{in})

$$R_a = \frac{1{\cdot}5}{2\pi \times f_{max} \times C_t}$$

$$L_c = 0{\cdot}67 \times C_t \times R_a^2$$

Relative gain at $f_{max} = 1{\cdot}5$.

" Shunt-Series "

($2 \times c_{out}$ should be adjusted to equal c_{in})

$$R_a = \frac{1{\cdot}8}{2\pi \times f_{max} \times C_t}$$

$$L_a = 0{\cdot}12 \times C_t \times R_a^2$$

$$L_c = 0{\cdot}52 \times C_t \times R_a^2$$

Relative gain at $f_{max} = 1{\cdot}8$.

The frequency response may also be improved in the usual manner by the use of negative feed-back. In general this allows for the use of higher values of anode load but the feed-back reduces the gain to a figure of the same order as described above. The response to wave shapes other than sine waves, however, is improved.

Wide-Band R.F. Amplifiers

These amplifiers generally employ either tuned anode, tuned grid or transformer coupling, but the tuned circuits are loaded with low values of parallel resistance. The tuning capacity is generally provided solely by the input and output capacities of the valves plus the stray capacities.

Two typical examples of circuit are shown in Fig. 8a and 8b. In Fig. 8a the inductance is in the grid circuit (L_g) the resistance loading being the anode load (R_a). In Fig. 8b the inductance is in the anode circuit the resistance loading being the grid leak (R_g). In both cases the tuning capacity is the total of $c_{in} + c_{out}$, and strays or C_t. The value of the coupling condenser C_o is such that its reactance is negligible at the mid-frequency point.

The band-width of either of these circuits for a response of 30 per cent. down at the band edges

$$2\Delta f = f_r \frac{\sqrt{L/C_t}}{R} \text{ c.p.s.}$$

where f_r is the mid-frequency resonance in Mc/s.

L = inductance of L_a or L_g in microhenrys.

C_t = total capacity in microfarads.

$2\Delta f$ = total band-width in cycles.

R = anode load (R_a) or grid leak (R_g).

The voltage gain at the mid-frequency is approximately that which would be obtained with an anode load of R by itself, *i.e.* $g_m \times R$.

When a wide-band R.F. or I.F. transformer is employed the voltage gain is approximately half the above value, if R is taken as the load resistance across the primary of the transformer.

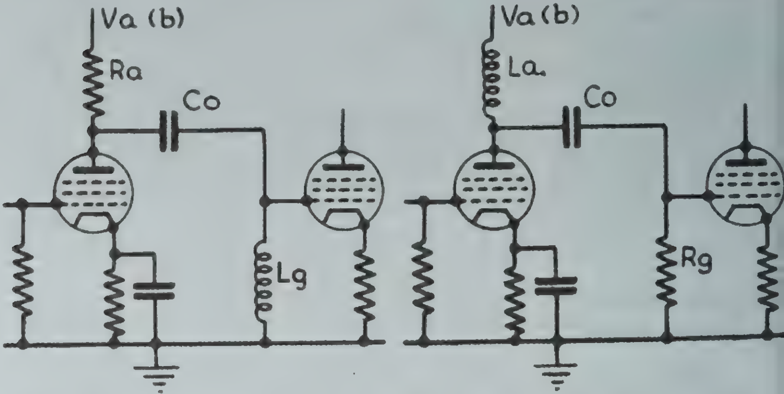


Fig. 8.
Two typical circuits for wide band R.F. amplifiers. (*a*) Using tuned grid coupling. (*b*) Using tuned anode coupling.

Use of Characteristic Curves

In order to determine the output voltage and distortion in wide-band amplifiers the same procedure is followed in all respects as described previously for resistance-capacity coupled or choke-capacity coupled amplifiers, but since the values of anode load resistor (R_a) are generally very small the anode voltage (V_a) is almost equal to the H.T. line voltage ($V_{a(b)}$). Hence for pentodes, dynamic curves or low screen voltage curves are not necessary.

CHAPTER 3 AUDIO FREQUENCY POWER AMPLIFIERS

The Five Point Method of Design

THE most essential characteristic curves for the design of the audio output stage are those showing the relationship between anode voltage and anode current at various No. 1 grid (control grid) potentials. In the case of pentodes and tetrodes the No. 2 grid (screen) potential is assumed constant and its current is often shown as well. Fig. 9a shows typical curves for a triode and Fig. 9b those of a pentode.

The "working point" is determined by finding the intersection of the vertical line through the anode supply voltage with the appropriate bias curve. The corresponding value of anode current may then be read off the anode current scale.

The load into which the valve works may be represented by a straight line, for resistive loads, passing through the working point, and having a slope equal to the reciprocal of the resistance in kilo-ohms. (assuming curves are plotted in volts and milliamperes).

In order to calculate power output and harmonic distortion, five special points must be determined. These are the intersections of the load line with the I_a V_a characteristics for certain grid potentials corresponding to the bias point, the bias point plus and minus half the peak grid swing and the bias point plus and minus the peak grid swing. Usually the bias point plus the peak grid swing will be equal to zero. These lines are shown drawn in on Fig. 10, together with a load line. Certain dimensions (capital letters for

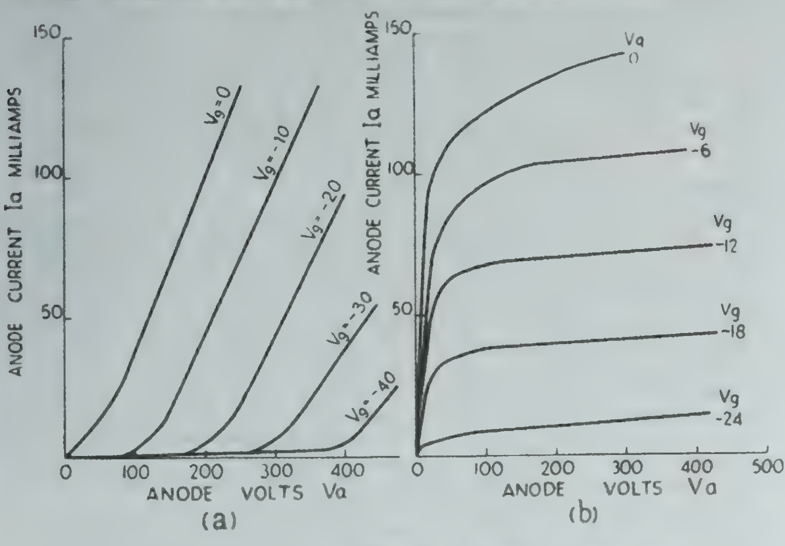
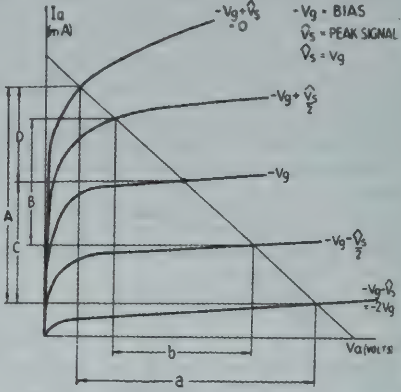


Fig. 9.
Typical I_a/V_g curves for an output valve. (*a*) Triode. (*b*) Pentode.

milliamperes, small letters for volts) can then be obtained, and from these, by use of the formula, all the information likely to be needed can be calculated. A little practice will soon enable a designer to select the most likely working point and the best load impedance for a given case. Slight variations of the values can be checked with the formula, to ensure that optimum conditions have been selected.



POWER OUTPUT

FUNDAMENTAL ONLY........ $\frac{Aa}{8}$ MILLIWATTS

WITH HARMONICS..... $\frac{(A+B)(a+b)}{18}$ MILLIWATTS

% SECOND HARMONIC.... $\frac{3(D-C)}{4(A+B)} \times 100\%$

% THIRD HARMONIC.... $\frac{A-2B}{2(A+B)} \times 100\%$

Fig. 10.
Load line drawn on typical pentode characteristics showing dimensions required to calculate output and harmonic distortion.

Class A. Single Valve

This case was illustrated in the foregoing section. It should be noted that the grid is never driven positive and, as no grid current flows, the previous stage need not supply power. The excursions of the anode current are equal either side of the no-signal value and no change of mean anode current should occur when a signal is applied. The no-signal dissipation will be given by the product of anode voltage and current at the working point, which must not exceed the makers' figures. The on-load dissipation will be this same figure minus the output power.

During a complete cycle of operation the anode potential traverses the whole of the load line, and at one point the anode reaches a potential nearly twice that of the supply voltage. The source of this additional voltage is the magnetic field in the core of the output transformer or choke, which, while the anode current is falling, collapses and induces in the winding a voltage which is additive to the supply voltage.

Were no choke or output transformer in use, the high tension supply voltage would have to be raised to the value indicated by the intersection of the load line with the anode voltage scale, to achieve the same working conditions.

Two Valves in Push-pull, Class A.

In this case it is necessary to construct composite characteristic curves representing the combination of the two valves, as shown in Fig. 11, and to apply the five point method as before. If the valves are assumed to have a fixed bias of –12v, and a signal of 12 volts peak per valve is applied, the corresponding pairs of grid voltages and the five relevant instants are shown in Table 1.

TABLE 1.

Instant	Grid Volts Valve No. 1.	Grid Volts Valve No. 2.
1	0	–24
2	–6	–18
3	–12	–12
4	–18	–6
5	–24	0

The characteristic curves of the two valves are combined by inverting one, so that the zero anode current lines are superimposed with the working anode voltages coincident. By algebraically adding the curves corresponding to the pairs of grid voltages given in Table 1 the composite curves may be constructed. For example, point A on the composite characteristic (drawn heavily), is obtained by adding the vertical dimensions x and y, y being negative and greater than x. The algebraic sum is $-(y - x)$ which being negative, is plotted below the horizontal axis.

The working point is the intersection of a vertical line through the working anode voltage with the centre line of the five composite curves. This will also lie on the horizontal axis, if the two valves are identical.

The load line may be drawn as before, the reciprocal of the slope representing the load applied to the composite generator. The five points are thus obtainable, but in extracting the necessary data to apply the formula it is better to make the measurements in linear units, and convert these to volts and milliamperes by the scale of the diagram, to avoid confusion arising over the negative values. The equivalent load on each of the individual valves can be obtained by drawing vertical lines through the five intersection points of the composite load line and the composite characteristics. Where these vertical lines cut the individual valve characteristic are points on the load line of the separate valves. (See A1, B1, C1, D1, E1, and A2, B2, C2, D2, E2 of Fig. 11). The slope of the individual load line is half that of the composite one indicating that the individual valve load is twice that of the composite generator. The anode-to-anode load will be four times that of the composite generator. This is shown another way in Fig. 12a, b and c, and may be taken as a general rule to be applied in all push-pull cases. If any operating condition other than pure Class A is examined, it will be found that the individual load lines are not straight.

Two Valves in Push-pull, Class AB1.

It is possible to attain a higher efficiency, still without positive grid drive, by over-biasing the valves, compared to the Class A bias. Under conditions of low signal level the anode current is reduced but increases as the signal increases. The degree of over-bias may vary from something barely distinguishable

from Class A to the "quiescent push-pull" condition when both valves, in the absence of a signal, are almost biased to cut off.

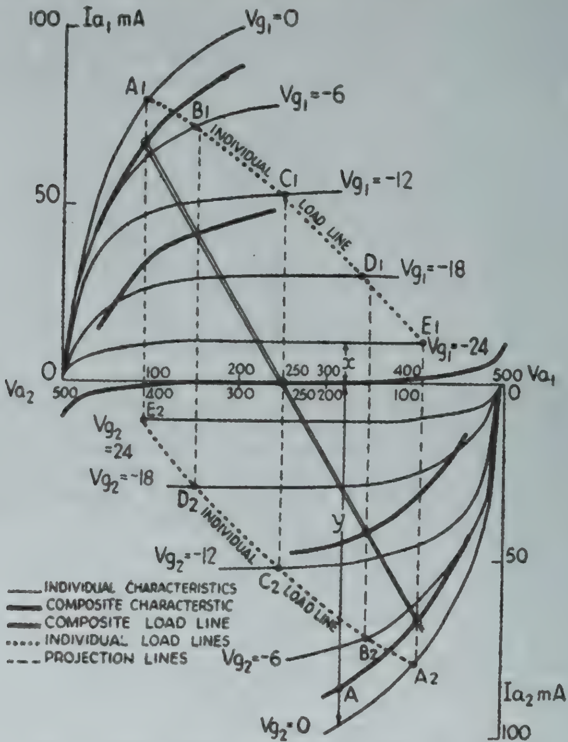


Fig. 11.
Composite diagram for a pair of pentodes in Class A, push-pull.

These conditions are illustrated for the case of a pair of imaginary triodes. One condition (Fig. 13a) assumes a bias of –22 and the other (Fig. 13b) of –30 volts. The corresponding "pairs of grid bias lines" are shown in Table 2.

The curved load lines of the individual valve characteristics will be noted, the curvature increasing with increase in standing bias. The composite characteristics, it will be observed, approach more closely the individual characteristics as the bias is raised, and in some applications this renders construction of the whole composite diagram unnecessary. Figures may be worked out by making the load line cut the working voltage at zero anode current and assuming symmetry about this point.

The dissipation under no-load conditions corresponds to the current and voltage values indicated by the centre point of the curved individual load line; the on-load dissipation however is harder to deduce. If apparatus is available the dissipation may, of course, be determined by measurement of

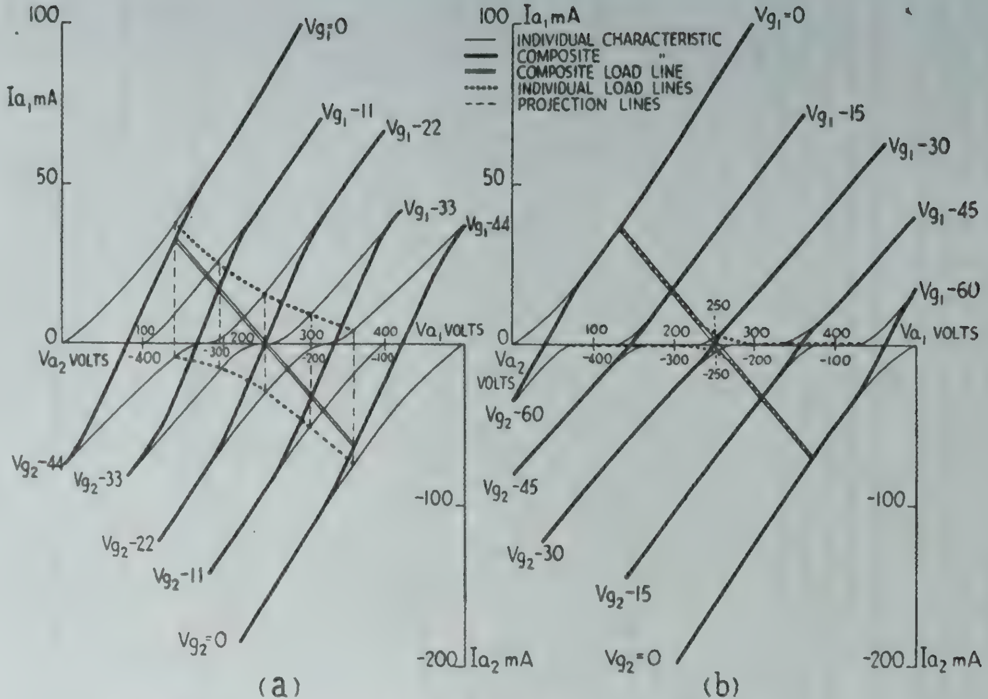


Fig. 13.
Composite diagrams for a pair of triodes in Class AB1 push-pull. (*a*) with a grid bias of — 22 volts. (*b*) with a grid bias of — 30 volts.

the input and calculation or measurement of the output, the total dissipation for both valves being the difference between these values. A calculated solution may be obtained as follows. First determine the instantaneous grid voltage at every 30° of one complete grid drive cycle of 360°. Second locate intersections of each of these grid bias lines with the curved load line and calculate the product of current and voltage corresponding to the intersections. Add the products and divide the sum by 12. For greater accuracy use 15° intervals for the grid drive voltage and divide by 24.

TABLE 2.

Instant.	Bias –22v. Signal 22 volts peak		Bias –30v. Signal 30 volts peak.	
	Valve 1.	Valve 2.	Valve 1.	Valve 2.
1	0	–44	0	–60
2	–11	–33	–15	–45
3	–22	–22	–30	–30
4	–33	–11	–45	–15
5	–44	0	–60	0

Two Valves in Push-pull, Class AB2.

Both Class AB2 and Class B differ from the previous examples because the peak grid drive voltage is greater than the standing bias and the valves are driven into the positive grid region. Because of the grid current, the previous stage, now known as the driver, has to supply power and itself has to be designed as a small output stage. Class AB2 is only a special case of Class AB1 in which the drive is taken beyond the zero grid voltage point to perhaps + 10 volts. The " pairs of bias " lines may then be, for example, as shown in Table 3. Providing the anode-current, anode-voltage curves for the positive values are available,

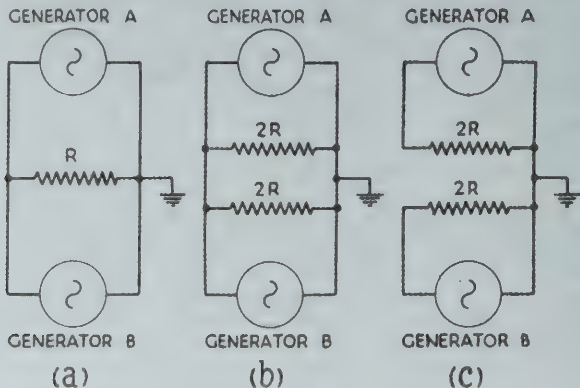


Fig. 12.
Illustrating the relationship between individual load, composite load and anode-to-anode load for a pair of valves Class A1 in push-pull. (*a*) Load on composite generator = *R*. (*b*) Two parallel resistors each 2 x *R*. Load identical to (*a*). (c) Two parallel resistors each 2 x *R* but fed from individual generators. Generator to generator impedance = 4 x *R*.

Table 3.

Instant.	Bias –30, Signal 40 volts peak.	
	Valve No. 1.	Valve No. 2.
1	+10	–70
2	–10	–50
3	–30	–30
4	–50	–10
5	–70	+10

the design of the stage proceeds exactly as before. At the instant of peak positive grid voltage, the sum of the anode and grid currents (screen current as well in the case of tetrodes or pentodes) must not exceed the permissible peak emission figure of the valve.

Two Valves in Push-pull, Class B.

It will be noted that in the Class AB2 case grid current flows for only a small fraction of each cycle, but some valves are designed to run into grid current for nearly a half cycle so that the push-pull pair draw grid current almost continually. This is the Class B condition. The valves, usually triodes, are designed to have low anode current consumption at zero or slightly negative voltage, and because the running condition is near to cut-off only one set of curves need be considered.

Fig. 14 shows the characteristics of a typical valve, the pairs of load lines being assumed to be as shown in Table 4. Three of the five points, marked A, B, C, are shown, the other two are assumed to be symmetrical about C.

Table 4.

Instant.	Bias –5, Signal 30v. peak.	
	Valve 1.	Valve 2.
1	+25	–35
2	+10	–20
3	–5	–5
4	–20	+10
5	–35	+25

Note also the grid current lines. At the plus 25 voltage instant the grid current is 23 milliamperes and the anode current 77 milliamperes. The peak demand is therefore 100 milliamperes.

The Driver Stage

In order to determine the operating conditions of the driver stage, two factors must be extracted from the grid current line corresponding to the peak positive drive voltage. They are the peak grid current, already noted to be 23 milliamperes, and the peak grid voltage from the bias point, which is not 25 volts but 25 – (– 5) *i.e.* 30 volts.

The total drive power, for two valves is equal to one half the peak drive volts (from bias point) multiplied by the peak grid current.

The load impedance of the grid circuit is equal to the peak grid voltage divided by the peak grid current flowing through one half of the secondary of the driver transformer at a time.

Driver transformer ratio (total primary to total secondary)

$$= \tfrac{1}{2}\sqrt{\frac{\text{Optimum load of driver} \times \text{peak grid current.}}{\text{Peak grid voltage}}}$$

This method is equally applicable to determining the driver stage conditions of a Class AB2 amplifier. The above information concerning the driver takes no account of the regulation of the driver stage, and for good quality this regulation must be good, *i.e.* the output voltage must not vary when the valve is required to deliver power to the next stage. This in turn demands a driver transformer of low D.C. resistance and a driver stage of low output impedance.

Output Impedance

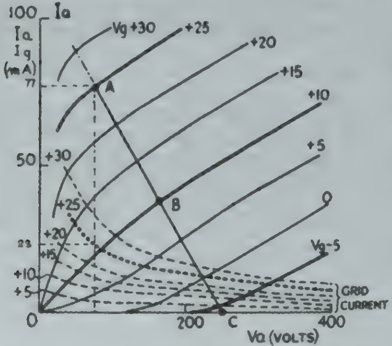


Fig. 14.
Anode current and grid current curves for one triode of a Class "B" push-pull pair.

The output valve may be regarded as an A.C. generator, of internal resistance r_a, working into a load represented by the output transformer and loudspeaker (or other load). This generator "sees" a load equivalent to the A.C. resistance of the loudspeaker speech-coil multiplied by the square of the step-down transformer ratio. The generator's internal impedance, or output impedance, is the reciprocal of the slope of the I_a/V_a curve at the working point, or the working point transferred to the individual characteristics in the case of push-pull pairs. The more nearly vertical the tangent the lower the impedance.

It can be seen, therefore, that pentodes have higher output impedances than triodes, and that Class AB and Class B stages are higher than Class A, since the working points of the AB and B stages are at lower current levels and, in consequence, are generally on portions of the *Va/Ia* curves that bend round towards the horizontal. In all cases, output impedances may be lowered by use of a suitable form of negative feed-back.

In order to ensure low output impedance and, therefore, improved regulation, Class AB2 and Class B driver stages are almost always push-pull triodes, but the best possible driver stage is a push-pull cathode follower, which may be regarded as a pair of triode valves with 100 per cent. negative feedback. This refinement is not usually necessary for output stages delivering 100 watts or less.

The output impedance of the audio stage is not usually of importance when the amplifier is used as a modulator, but if it is to be used with a loud speaker, a low output impedance will be found to have a marked effect in damping loud speaker resonances and generally assures a more constant output across the frequency spectrum. The object should be to make the ratio of load impedance to output impedance as high as possible with a minimum of four to one for the best quality. Push-pull triodes Class A or Class AB1 with some negative feedback have never been surpassed for high quality reproduction, nor does it seem likely that they will be, although pentodes, if sufficient feedback is used, can be made as good.

CHAPTER 4 RADIO FREQUENCY POWER AMPLIFIERS, FREQUENCY MULTIPLIERS AND OSCILLATORS

THE following additional symbols are used in this Chapter.

$v_{a\,(pk)}$	Peak anode voltage.
$v_{g\,(pk)}$	Peak grid voltage.
$v_{g\,(max)}$	Maximum positive grid voltage.
$v_{a\,(min)}$	Minimum anode voltage.
$i_{a\,(pk)}$	Peak anode current.
$i_{g\,(pk)}$	Peak grid current.
$'$	Indicates a value at the modulation crest, *e.g.* V_a'.
2θ	The angle during which anode current flows.

$$K_1 = \frac{1}{1 - \cos\theta} \qquad \text{(see Fig. 18)}$$

$$K_2 = K_1 - 1$$

$$K_3 = \frac{i_{a\,(pk)}}{I_a} \qquad \text{(see Fig. 18)}$$

$$K_4 = \frac{W_{out}}{I_a \times v_{a\,(pk)}} \qquad \text{(see Fig. 19)}$$

R.F. POWER AMPLIFIERS

In an R.F. power amplifier, the anode current consists of a series of pulses which may occupy between 40° and 180° of each complete R.F. cycle of 360°, depending on the characteristics of the valve and the purpose for which it is used. The operating conditions can rarely be calculated precisely, but an accuracy is obtainable which is sufficient to permit of intelligent transmitter adjustment. For all power amplifier design, valve characteristic curves extending into the positive grid region showing anode and grid currents are required. Curves of a typical triode valve are shown in Figs. 16 and 17 and these are used in the worked examples.

Fundamental Relationships

Fig. 15a shows a skeleton circuit for a triode power amplifier and Fig. 15b the relationship between the various voltages and currents in this circuit. It should be particularly noted that the peak value of anode current is drawn at an instant when the anode voltage has reached a low value and the grid voltage has reached a maximum positive value. In designing an amplifier this point has to be selected on the I_a/V_a and I_g/V_a curves. The anode voltage should never be allowed to fall below the value of the positive grid voltage as this results in excessive driving power being required

and may damage the grid of the valve; on the other hand, the efficiency will be greater when the anode voltage swings down to a very low value. With practice, it becomes easy to select a suitable working point by inspection of the valve characteristics; until skill is obtained, the value of minimum anode voltage may be taken as 20 per cent. of the D.C. anode voltage for initial calculations.

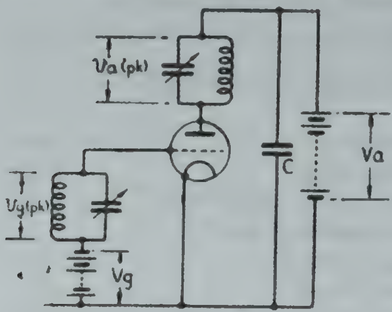


Fig. 15a.
Basic circuit of Class C amplifier.

Fig. 15c shows the conditions when anode modulation is applied to the amplifier. Modulation is accomplished by varying the anode voltage; at the crest of modulation the anode voltage is doubled and the peak R.F. anode voltage is also doubled. Normally the peak value of the R.F. grid voltage is unchanged since no modulation is applied to the driver stage; since the peak anode current is increased at the modulation crest, a greater value of positive grid voltage is required. This is achieved by reducing the value of D.C. grid bias or occasionally by applying some modulation to the driver.

Fig. 15d shows the conditions when grid modulation is applied. At the crest of modulation the anode voltage will swing down to a value approaching the positive grid voltage similar to the conditions for Class C telegraphy but since the bias with no modulation is increased the positive grid voltage, peak anode current and the anode voltage swing, are all reduced. Modulation is accomplished by effectively varying this bias. The peak grid voltage supplied from the driver must not vary during the modulation cycle in spite of the variation in driving power that is required.

When designing a power amplifier, it is necessary to decide during what part of each cycle the anode current shall flow. In general short angles of flow produce high efficiencies but demand greater driving power and take greater peak emission from the cathode of the valve than do large angles. This may result in reduced valve life. The angle can be chosen for each individual case bearing those factors in mind; in the design

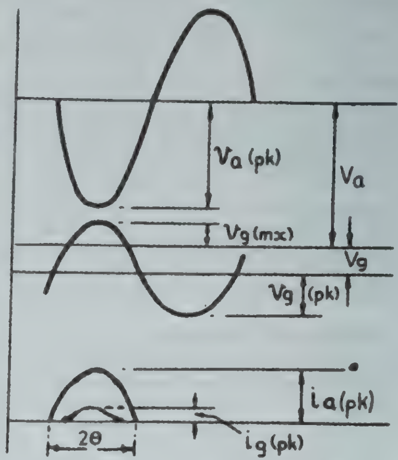


Fig. 15b.
Phase relationship in an unmodulated Class C amplifier.

data that follows, typical angles have been chosen and these may be adopted for most normal applications. It should be noted that the angle "θ" used in the design curves of Figs. 18 and 19 is one-half of the angle of flow; this is done to simplify the formulæ.

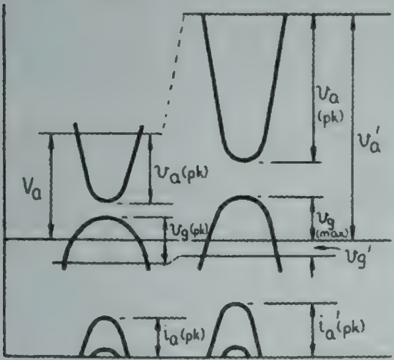


Fig. 15c.
Phase relationship at the carrier and modulation crest for an anode modulated Class C amplifier.

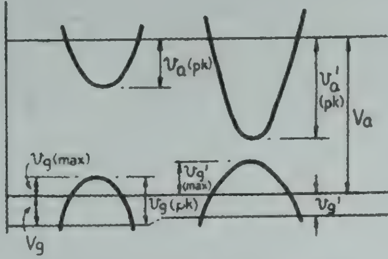


Fig. 15d.
Phase relationship at the carrier and modulation crest for a grid modulated Class C amplifier.

The following sections indicate the design procedure for Class C amplifiers for telegraphy, anode modulation, grid modulation and for frequency multiplication. Each section concludes with a summary in which the various calculated values are rounded off so as to indicate practical values.

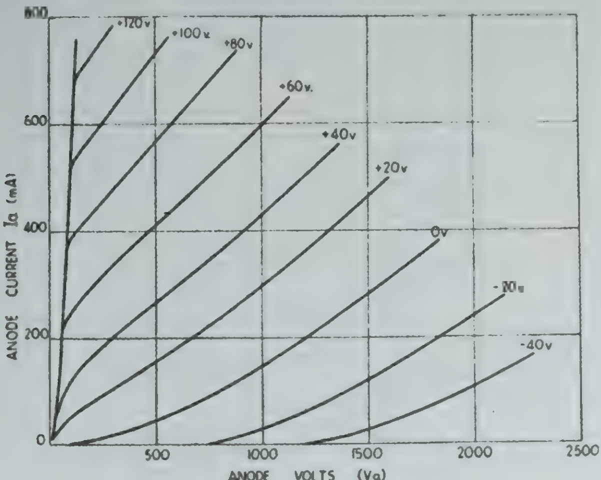


Fig. 16.
I_a/V_a characteristics for a typical transmitting triode.

Class C. Telegraphy

It is first necessary to decide the values of V_a and I_a, the product of which is the anode input. This is, of course, limited by the power that the station is licensed to use. Consider the valve whose characteristics are shown in Figs. 16 and 17; it is a triode of amplification factor (μ) 30 whose maximum ratings are $V_a = 1{,}200$ volts, $I_a = 125$mA, anode dissipation (W_a) = 40 watts. The maximum input will be: $1{,}200 \times 0{\cdot}125 = 150$ watts which is the usual maximum for an amateur station.

The design equations are:—

$$i_{a(pk)} = K_3 \times I_a \quad \ldots\ldots (1)$$

$$v_{a(pk)} = V_a - v_{a(min)} \quad \ldots\ldots (2)$$

$$W_{out} = K_4 \times I_a \times v_{a(pk)} \quad \ldots\ldots (3)$$

$$Vg = K_2\left\{-v_{g(max)} + \frac{v_{a(pk)}}{\mu}\right\} - K_1 \times \frac{V_a}{\mu} \quad \ldots\ldots (4)$$

$$Za = \frac{v_{a(pk)}}{2 \times I_a \times K_4} \quad \ldots\ldots (5)$$

A normal angle of flow for Class C telegraphy is 130°, so that $\theta = 65°$. For this angle, the four design factors can be obtained from Figs. 18 and 19 and are as follows:—

$K_1 = 1{\cdot}75$; $K_2 = 0{\cdot}75$; $K_3 = 4{\cdot}3$; $K_4 = 0{\cdot}88$.

From equation (1) $i_{a(pk)} = 4{\cdot}3 \times 125 = 540$ mA.

This is drawn at a low value of anode voltage such as 200 volts requiring a positive grid excursion of + 95 volts and producing a peak grid current of 110 mA. These figures are found from the characteristic curves.

From equation (2) $v_{a(pk)} = 1{,}200 - 200 = 1{,}000$ volts.

From equation (3) $W_{out} = 0{\cdot}88 \times 0{\cdot}125 \times 1{,}000 = 110$ watts.

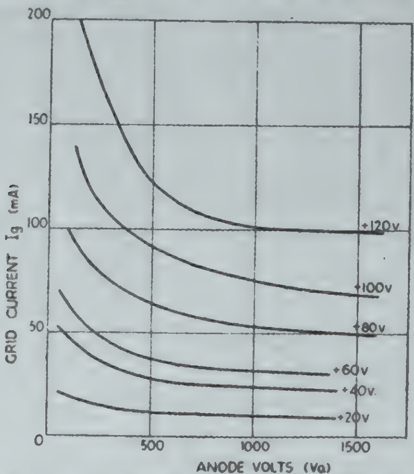


Fig. 17.—Corresponding I_g/V_a characteristics for the transmitting triode shown in Fig. 16.

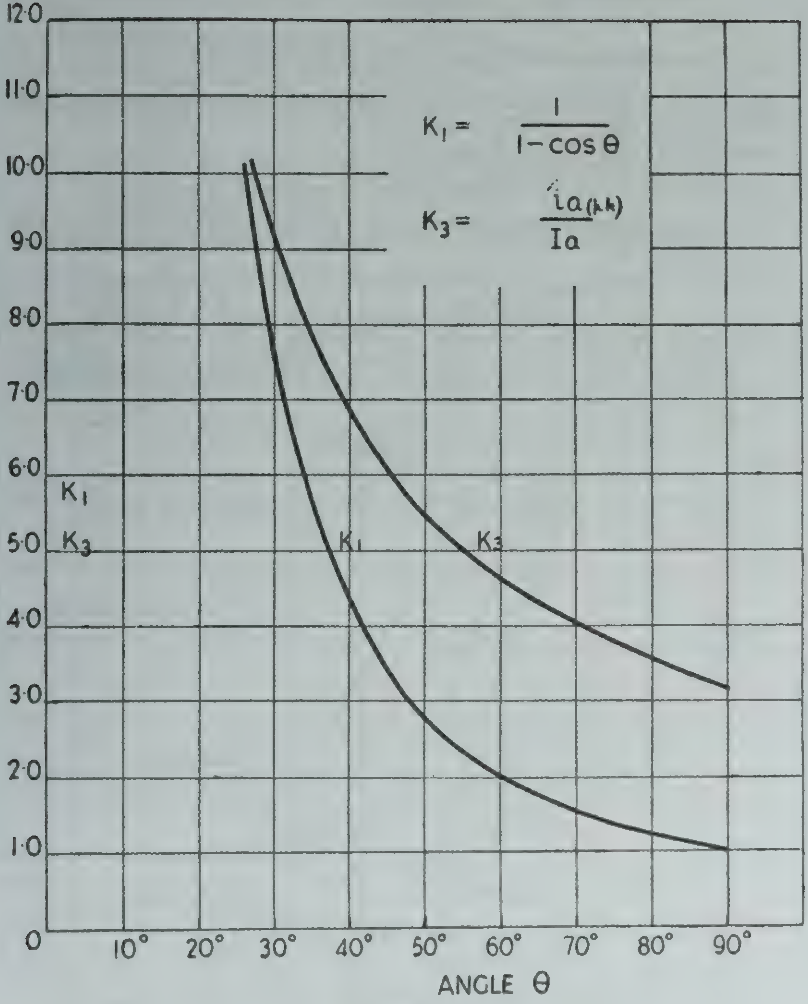


Fig. 18.
Values of the constants K_1 and K_3 for various angles of flow.

Subtracting this from the anode input, it is found that the anode dissipation is not exceeded.

From equation (4)

$$Vg = 0{\cdot}75 \times \left(-95 + \frac{1{,}000}{30}\right) - 1{\cdot}75 \times \frac{1{,}200}{30} = -117 \text{ volts.}$$

The peak grid swing $v_{g\,(pk)} = 117 + 95 = 212$ volts.

The ratio $\dfrac{V_g}{v_{g\,(pk)}} = \dfrac{117}{212} = 0{\cdot}55$ and referring this to Fig. 20 we find the ratio of peak to D.C. grid current is $7{\cdot}1$

$$\text{whence } I_g = \frac{110}{7{\cdot}1} = 16 \text{ mA.}$$

The driving power required is given by the product of D.C. grid current and peak grid swing $= 0{\cdot}016 \times 212 = 3{\cdot}5$ watts. The driver stage should pro-

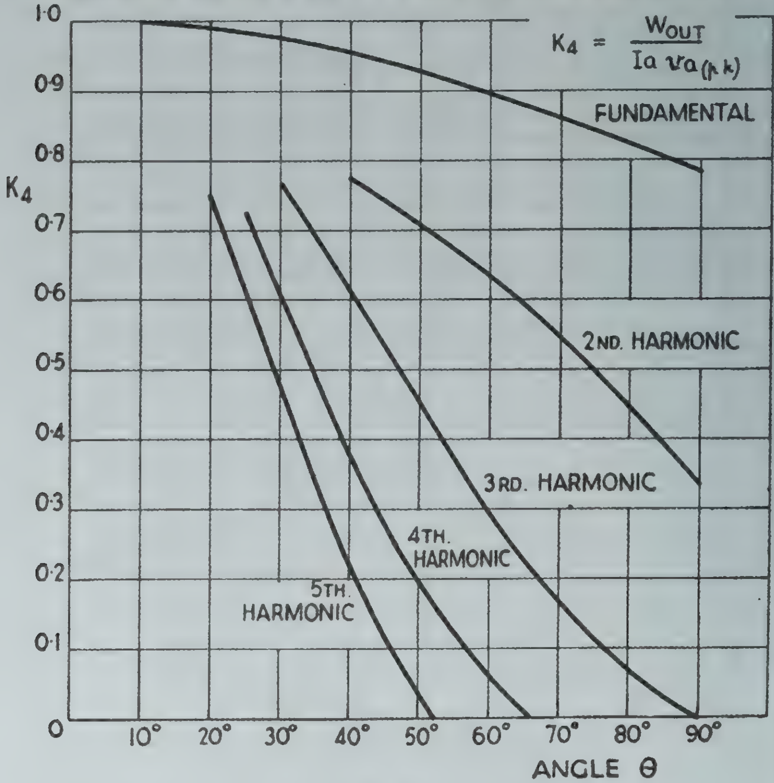


Fig. 19.
Values of the constant K_4 for various angles of flow utilising the fundamental and various harmonics.

duce considerably more power than this to allow for losses in the coupling system.

From equation (5) $Z_a = \dfrac{1{,}000}{2 \times 0{\cdot}125 \times 0{\cdot}88} = 4{,}550$ ohms.

Summarising the design—

(1) the anode supply will be 1,200 volts at 125 mA;
(2) the grid bias will be −120 volts;
(3) the previous stage should give an output of about 6 watts and the inter-stage coupling should produce 215 volts peak at the amplifier grid;
(4) the output will be 110 watts, less the loss in the tuned circuit.
(5) the anode tuned circuit, when loaded by the aerial should present a load of 4,500 ohms.

The design of practical tuned circuits is dealt with in a later section.

Anode-modulated Class C. Telephony

Valve makers usually quote ratings for their products under anode-modulated conditions but if only Class C telegraphy ratings are known, it is sufficient to reduce the anode voltage and current to 80 per cent., and the anode dissipation to 65 per cent. of the maximum values allowed for Class C telegraphy.

For our example, the ratings become:—

$V_a = 1{,}200 \times 0{\cdot}8 = 960$ volts (say 1,000 volts)

$I_a = 125 \times 0{\cdot}8 = 100$ mA

Anode dissipation (W_a) $= 40 \times 0{\cdot}65 = 26$ watts.

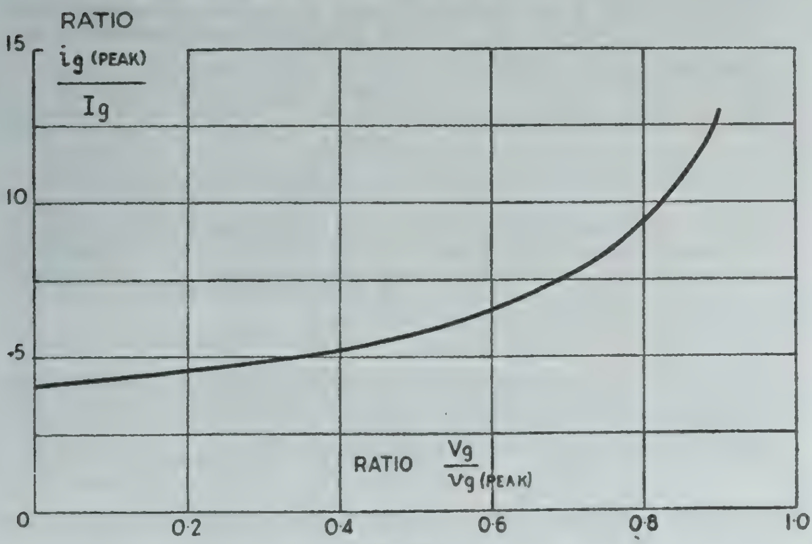


Fig. 20.
A curve to determine the value of D.C. grid current as based on the analysis of a squared sine wave.

The early part of the design follows along exactly the same lines as the calculation for Class C telegraphy.

A normal angle of anode current flow at the carrier is 120° so that $\theta = 60°$ and the design factors are $K_1 = 2{\cdot}0$; $K_2 = 1{\cdot}0$; $K_3 = 4{\cdot}6$; $K_4 = 0{\cdot}9$.

From equation (1) $i_{a\,(pk)} = 4{\cdot}6 \times 100 = 460$ mA.

The value of $v_{a\,min}$ (120 volts) is selected from the curves whence

$$v_{g\,max} = +\,75 \text{ volts}$$

and the peak grid current = 85 mA.

From equation (2) $v_{a\,(pk)} = 1{,}000 - 120 = 880$ volts.

From equation (3) $W_{out} = 0{\cdot}9 \times 0{\cdot}100 \times 880 = 80$ watts.

From equation (4) $V_g = 1{\cdot}0\left(-75 + \frac{800}{30}\right) - 2{\cdot}0 \times \frac{1{,}000}{30} \simeq -112$ volts.

The peak grid swing $v_{g\,(pk)} = 112 + 75 = 187$ volts.

The ratio $\frac{V_g}{v_{g\,(pk)}} = \frac{112}{187} = 0{\cdot}6$ and from Fig. 20 we find the ratio of peak to D.C. grid current $= 6{\cdot}5$ whence $I_g = \frac{85}{6{\cdot}5} = 13$ mA. The driving power $= 0{\cdot}013 \times 112 = 1{\cdot}5$ watts.

From equation (5) $Z_a = \frac{880}{2 \times 0{\cdot}100 \times 0{\cdot}90} = 4{,}900$ ohms.

In order to ensure that full modulation is possible, the conditions at the modulation crest should be checked.

A typical angle of flow at the crest is 180° so that $\theta = 90°$ and the design factors are $K_3' = 3{\cdot}14$; $K_4' = 0{\cdot}785$. The anode voltage at the crest will be doubled, *i.e.* $V_a' = 2{,}000$ volts.

The peak anode current $i_{a\,(pk)}' = \dfrac{2 \times W_{out} \times K_3'}{v_{a\,(pk)} \times K_4'}$ ....(6)

$$i_{a\,(pk)}' = \frac{2 \times 80 \times 3{\cdot}14}{880 \times 0{\cdot}785} = 720 \text{ mA.}$$

It should be noted that this is about the highest peak current that could have been demanded. If a higher value had been demanded, it would have been necessary to repeat the design using a lower value of D.C. anode current.

The anode voltage at which the peak current is drawn must be twice that used at the carrier if modulation is to be linear.

$$v_{a\,(min)}' = 2 \times 120 = 240 \text{ volts.}$$

The positive grid voltage required $= +\,125$ volts and the resulting peak grid current = 180 mA. Since the peak grid voltage does not vary over the modulation cycle, the grid bias must change from the carrier value so as to provide the positive grid voltage required at the crest.

$$V_g' = 187 - 125 = 62 \text{ volts.}$$

The ratio $\frac{V_g}{v_{g\,(pk)}} = \frac{62}{187} = 0{\cdot}33$ whence Fig. 20 shows a ratio of peak to D.C. grid current of $5{\cdot}0$. Hence $I_g' = \frac{180}{5{\cdot}0} = 36$ mA.

In some cases, it will be found that the grid current at the crest is less than that at the carrier. In such cases, the change in bias is easy to obtain, since a value of grid leak can be selected such that the required voltage change is produced by the calculated current change. In other cases, of which our example is typical, the grid current increases towards the crest so that a grid leak tends to shift the bias in the wrong direction. One solution to this problem is to apply a little modulation to the previous stage. This ensures that there is sufficient positive grid voltage to provide the crest anode current even though the bias is higher than the calculated value. It should be remembered that grid current calculations are subject to quite considerable errors due to the steep rise of grid current at low values of anode voltage. The calculations should be regarded as no more than indicating the type of experimental adjustments that are most likely to produce linear modulation.

The impedance offered to the modulator is given by $\dfrac{V_a}{I_a} = \dfrac{1{,}000}{0 \cdot 100} = 10{,}000$ ohms and the modulating power required is half the anode input to the R.F. amplifier $= 0 \cdot 5 \times 1{,}000 \times 0 \cdot 100 = 50$ watts.

Summarising the design:

(1) the anode supply will be 1,000 volts at 100 mA.
(2) the grid bias at the carrier will be — 110 volts.
(3) the previous stage should give an output of about 5 watts (a liberal allowance is desirable) and the inter-stage coupling should produce 185 volts peak at the amplifier grid. This will produce a D.C. grid current of about 13 mA.
(4) the power output will be 80 watts, less the loss in the tuned circuit.
(5) the modulator should deliver 50 watts into a load of 10,000 ohms.

Grid-modulated Class C Telephony

Since the conditions at the modulation crest are similar to those for Class C telegraphy, they will be calculated first. A typical angle of anode current flow is 180° so that $\theta = 90°$ and the design factors are $K_1 = 1 \cdot 0$; $K_2 = 0$; $K_3 = 3 \cdot 14$; $K_4 = 0 \cdot 785$.

The maximum anode voltage and dissipation are the same as for Class C telegraphy, but as the efficiency without modulation is essentially low, the permissible anode input is considerably reduced.

At the crest of modulation, the efficiency will be about 65 per cent. and from this the anode input can be assessed.

$$\text{Crest anode input} = \frac{40}{1 - 0 \cdot 65} = 114 \text{ watts.}$$

If $V_a = 1{,}200$ volts, the anode current $I_a{}' = \dfrac{114 \times 10^3}{1{,}200} = 95$ mA.

From equation (1) $i_{a\,(pk)}{}' = 3 \cdot 14 \times 95 = 300$ mA.

Let this current be drawn at $v_{a\,(min)}{}' = 150$ volts, requiring a positive grid excursion $v_{g\,(max)}{}' = +\,65$ volts and producing a peak grid current of 65 mA.

From equation (2) $v_{a\,(pk)}{}' = 1{,}200 - 150 = 1{,}050$ volts.

From equation (3) $W_{out}{}' = 0 \cdot 785 \times 0 \cdot 095 \times 1{,}050 = 78 \cdot 5$ watts.

From equation (4) $V_g{}' = \dfrac{-\,V_a}{\mu}$ (since $K_2 =$ zero and $K_1 = 1$)

$$= \frac{-1{,}200}{30} = -40 \text{ volts}$$

$v_{g\,(pk)}' = 40 + 65 = 105$ volts whence the ratio $\frac{V_g}{v_{g\,(pk)}} = \frac{40}{105} = 0{\cdot}38$ which when referred to Fig. 20 shows that the ratio of peak to D.C. grid current $= 5{\cdot}3$. Hence $I_g' = \frac{65}{5{\cdot}3} = 12{\cdot}5$ mA.

The driving power at the modulation crest $= 0{\cdot}0125 \times 105 = 1{\cdot}3$ watts. The driving power varies during the modulation cycle and is greatest at the crest.

Now it is necessary to determine the carrier conditions. A typical value for $\theta = 60°$ whence $K_1 = 2{\cdot}0$; $K_2 = 1{\cdot}0$; $K_3 = 4{\cdot}6$; $K_4 = 0{\cdot}9$.

The anode current $I_a = \frac{W_{out}'}{2 \times v_{a\,(pk)}' \times K_4}$ ....(7)

$$= \frac{78{\cdot}5}{2 \times 1{,}050 \times 0{\cdot}9} = 41{\cdot}5\text{mA.}$$

From equation (1) $i_{a\,(pk)} = 4{\cdot}6 \times 41{\cdot}5 = 190$ mA.

For modulation to be linear, the value of peak anode voltage must be exactly half of the crest value $= \frac{1{,}050}{2} = 525$ volts.

Hence $v_{a\,(min)} = 1{,}200 - 525 = 675$ volts.

Locating 190 mA and 675 volts on the I_a/V_a curves, we find a grid voltage $v_{g\,(max)} = +18$ volts producing a peak grid current of 10 mA. The peak grid voltage remains unchanged during the modulation cycle so that the grid bias at the carrier $= 105 - 18 = 87$ volts.

The bias at the crest was -40 volts and the difference between these two values is the peak modulating voltage required.

$$= 87 - 40 = 47 \text{ volts peak.}$$

The power required from the modulator equals half the product of the peak modulating voltage and the crest grid current

$$= \frac{47 \times 0{\cdot}0125}{2} = 0{\cdot}3 \text{ watts.}$$

The impedance offered to the modulator $= \frac{47^2}{2 \times 0{\cdot}3} = 3{,}700$ ohms.

The power output will be one quarter of that at the crest

$$= \frac{78{\cdot}5}{4} = 19{\cdot}6, \text{ say } 20 \text{ watts.}$$

The anode input $= 1{,}200 \times {\cdot}0415 = 50$ watts and the anode dissipation is $50 - 20 = 30$ watts. This is within our limit of 40 watts and it would now be possible to repeat the design using a higher anode input in an attempt to exploit fully the permissible anode dissipation.

The ratio $\frac{V_g}{v_{g\,(pk)}} = \frac{87}{105} = 0{\cdot}82$ and from Fig. 20 the ratio of peak to D.C. grid current is $9{\cdot}5$; hence $I_g = \frac{10}{9{\cdot}5} = 1$ mA (approx.)

From equation (5) $Z_a = \frac{525}{2 \times 0{\cdot}0415 \times 0{\cdot}9} = 7{,}000$ ohms.

Summarising the design:

(1) the anode supply will be 1,200 volts at 40 mA.
(2) the grid bias will be − 90 volts.
(3) the previous stage should give an output of about 2·5 watts and the inter-stage coupling should produce 110 volts peak at the amplifier grid. This will result in about 1 mA of grid current.
(4) the power output will be 20 watts less the circuit loss.
(5) the modulator should be designed to produce 50 volts peak into a load of 4,000 ohms. It should give a power output of about 0·5 watts.

FREQUENCY MULTIPLIERS

Frequency multipliers are designed in the same way as amplifiers for Class C telegraphy, but since the anode circuit is tuned to a multiple of the grid circuit frequency some of the formulæ are modified. Fig. 19 shows values of K_4 for harmonics up to the fifth; calculations of higher order harmonics are scarcely accurate enough to be worth while. The angle of anode current flow has to be chosen with some care; small angles which result in high efficiencies often require excessive values of peak anode currents and driving power. In order to estimate the permissible anode current, it is necessary to have some idea of the anode efficiency that will be obtained. The following table gives typical efficiencies and angles of flow.

Harmonic	2	3	4	5
Efficiency %	50	40	30	20
Angle θ	50°	40°	30°	25°

Let us consider the 40 watt triode used as a quadrupler. If the efficiency is taken as 30 per cent. the anode input $= \frac{40}{1 - 0{\cdot}3} = 57$ watts and if we assume $V_a = 1{,}000$ volts, then $I_a = \frac{57}{1{,}000} = 57$ mA. Taking $\theta = 30°$, $K_1 = 7{\cdot}7$; $K_2 = 6{\cdot}7$; $K_3 = 9{\cdot}2$; $K_4 = 0{\cdot}6$.

From equation (1) $i_{a\,(pk)} = 9{\cdot}2 \times 57 = 525$ mA.

The corresponding value of anode voltage is selected from the characteristic curves as 200 volts, requiring a positive grid voltage of + 94 volts and producing a peak grid current of 105 mA.

From equation (2) $v_{a\,(pk)} = 1{,}000 - 200 = 800$ volts.

From equation (3) $W_{out} = 0{\cdot}6 \times 0{\cdot}057 \times 800 = 27$ watts.

The equation for finding the grid bias is modified and becomes

$$V_g = -\frac{K_1}{\mu}\left(V_a - v_{a\,(pk)} \cos n\theta\right) - K_2 \times v_{g\,(max)} \quad \ldots\ldots\ldots(8)$$

where n is the number of the harmonic. Remember that if nθ is greater than 90°, the angle to be looked up in the cosine tables is (180° − nθ) and the result is negative. This is the case in our example where $n\theta = 4 \times 30° = 120°$ and $\cos 120° = -\cos(180° - 120°) = -\cos 60° = -0{\cdot}5$.

$$V_g = -\frac{7 \cdot 7}{30}\left(1{,}000 - 800\,(-0 \cdot 5)\right) - 6 \cdot 7 \times 94 = -990 \text{ volts.}$$

It is at once obvious that a grid voltage as high as the anode voltage is absurd and apart from requiring very great driving power would probably damage the valve. Therefore, the calculation must be repeated using a larger angle of flow. Let us try $\theta = 40°$ when $K_1 = 4 \cdot 4$; $K_2 = 3 \cdot 4$; $K_3 = 6 \cdot 9$; $K_4 = 0 \cdot 375$.

From equation (1) $i_{a\,(pk)} = 57 \times 6 \cdot 9 = 400$ mA

Let this occur at $v_{a\,(min)} = 125$ volts and $v_{g\,(max)} = +\,80$ volts

From equation (2) $v_{a\,(pk)} = 1{,}000 - 125 = 875$ volts

From equation (3) $W_{out} = 0 \cdot 375 \times 0 \cdot 057 \times 875 = 18 \cdot 5$ watts

From equation (8) $V_g = -\frac{4 \cdot 4}{30}\left(1{,}000 - 875\,(-\,\cdot 94)\right) - 3 \cdot 4 \times 80$
$= -\,540$ volts.

Even this might be considered an excessive value but let us accept it for the purpose of illustrating the design. The peak grid swing $v_{g\,(pk)} = 540 + 80 = 620$ volts. The ratio $\frac{V_g}{v_{g\,(pk)}} = \frac{540}{620} = 0 \cdot 87$ whence the ratio of peak to D.C. grid current is $11 \cdot 4$ (Fig. 20). The peak grid current is found from the characteristic curves to be 92 mA whence $I_g = \frac{92}{11 \cdot 4} = 8$ mA. The driving power $= 0 \cdot 008 \times 620 = 5$ watts.

From equation (5) $Z_a = \frac{875}{2 \times 0 \cdot 057 \times 0 \cdot 375} = 20{,}000$ ohms.

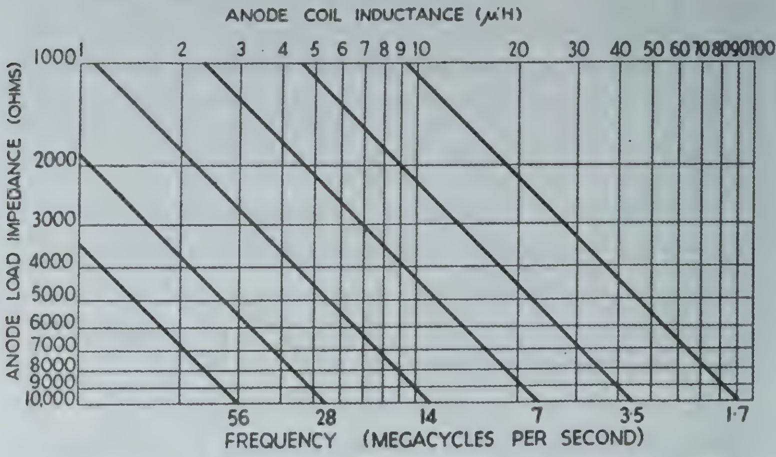


Fig. 21.
A chart from which can be read the values of anode tank coil inductance to suit the calculated anode load impedance for the various amateur bands.

Summarising the design:

(1) the anode supply will be 1,000 volts at 60 mA;
(2) the grid bias will be -540 volts;
(3) the previous stage should give an output of about 10 watts and the inter-stage coupling should produce 620 volts peak at the quadrupler grid. The grid current will be about 8 mA;
(4) the output will be 18 watts less the loss in the tuned circuit;
(5) the anode tuned circuit when loaded should present a load of 20,000 ohms.

Application of the Design Methods to Pentodes and Beam Tetrodes

All the design data given is for triode valves but the method is equally applicable to pentodes and beam tetrodes. Since the anode current depends largely on the screen voltage, the formulæ for calculating the grid bias are modified. Equations (4) and (8) become

$$V_g = -K_2 \times v_{g\,(max)} - K_1 \times \frac{V_{g2}}{\mu} \quad \ldots\ldots\ldots\ldots(9)$$

where v_{g2} is the D.C. screen voltage, and μ is the triode amplification factor of the valve. If the curves of screen current are available, the D.C. screen current can be calculated by taking the peak screen current (at $v_{a\,(min)}$ and $v_{g\,(max)}$) and dividing by K_3 to get D.C. values.

CIRCUIT DESIGN FOR R.F. POWER AMPLIFIERS AND FREQUENCY MULTIPLIERS

The Anode Tuned Circuit

Each of the design methods shown yields a figure for the load impedance Z_a which should be presented to the valve to obtain the calculated performance. This figure is used to design the tuned circuit.

Let Q_L = the Q of the circuit when unloaded

and Q_o = the Q of the circuit when loaded

Since $Z_a = Q_o \omega L$ it follows that if we decide on a value for Q_o, the reactance of the tuned circuit is known and the required inductance value can be determined for any frequency. It is usually accepted that $Q_o = 10$ gives a reasonable compromise between tuned circuit loss and harmonic content of the output so that $\omega L = \frac{Z_a}{10}$.

The chart of Fig. 21 simplifies the calculation and shows the inductance required for load impedances up to 10,000 ohms for the various amateur bands. This is the inductance required when the anode is connected to one end of the circuit and the other end is earthy. If the centre of the coil is made earthy as in a push-pull amplifier or a neutralised single ended amplifier, the inductance value should be increased by four times in order that the valve shall " see " the same impedance. As an example, take our Class C telegraphy design which calls for an anode load impedance of 4,500 ohms. For the 3·5 Mc/s. band, Fig. 21, shows an inductance of 20μH. Since our valve is a triode, it will probably be used in a neutralised circuit in which the centre of the coil is earthy so that an inductance of 80 μH is required. An Abac shows

a corresponding value for the tuning condenser of 24 pF; it should be remembered that this includes the anode-earth capacitance of the valve together with the stray capacitance.

At the higher frequencies and in the case of frequency multipliers which require high values of load impedance, it will often be found that the inductance required is so large that it will not tune to the desired frequency even with the smallest possible value of tuning condenser. In these cases, the largest practicable coil is used and the value of Q_o is increased so as to present the correct load impedance. This increases the tuned circuit losses, but cannot be avoided. For instance, our frequency quadrupler required a load impedance of 20,000 ohms. If the output is at 28 Mc/s. the inductance required is 22μH (ten times the value for 2,000 ohms in Fig. 21). The corresponding value of capacitance is 1·5 pF whereas the minimum practical value is probably nearer to 10pF. This allows an inductance of 3·2μH which represents a reactance of 600 ohms at 28 Mc/s.

Now $Q_o = \dfrac{Z_a}{\omega_L} = \dfrac{20{,}000}{600} = 33.$

The efficiency of a tuned circuit is given by $\dfrac{Q_L - Q_o}{Q_L}$. Thus if a tuned circuit has an unloaded Q of 100 and the circuit is loaded so that $Q_o = 10$, the efficiency is $\dfrac{100 - 10}{100} = 0{\cdot}9$. This means that 90 per cent. of the power output from the valve will be passed on to the aerial so that our Class C telegraphy design would deliver $110 \times 0{\cdot}9 = 99$ watts to the aerial. The quadrupler would have a tuned circuit efficiency of only $\dfrac{100 - 33}{100} = 0{\cdot}66$ so that only $18{\cdot}5 \times 0{\cdot}66 = 12$ watts will be available for the next stage.

Of course, the amateur station is rarely equipped with means of measuring Q, but the examples give an idea of the losses to be expected. Abacs exist from which the Q can be calculated with fair accuracy.

Grid Bias

Grid bias for Class C amplifiers may be obtained from a fixed supply, a cathode resistor or a grid leak or from a combination of these methods. Grid leak bias alone should normally be avoided because the bias disappears if the drive fails; this usually results in a serious increase in anode dissipation of the valve. Some grid leak bias is usually desirable for anode-modulated amplifiers as it tends to improve the modulation characteristics by allowing the grid bias to vary over the modulation cycle. Grid leak bias must not be used for grid-modulated amplifiers since a variation in bias would involve serious distortion.

Decoupling Condensers

The condenser which connects the anode tuned circuit to earth (C in Fig. 15a) must be rated to carry the R.F. current flowing in this circuit. This is usually quite small in the case of triodes, but may be considerable when pentodes are used. In a pentode, the capacitance between anode and suppressor grid (screen grid in the case of a tetrode) is often quite large and constitutes the major part of the anode-earth capacitance. Since the suppressor and screen grids are at R.F. earth potential, the R.F. anode voltage is applied

directly across the capacitance between the anode and these electrodes. If this is 20 pF, at 28 Mc/s. it represents a reactance of only 280 ohms and if the R.F. anode volts are 1,000 volts, a current of $\frac{1,000}{280} = 4$ amps approx. will flow in the condenser C. The cheapest and often the best form of condenser for such a position consists of two discs of brass about 2" diameter clamped on either side of the metal baseplate with mica insulation. The inner faces of the discs should be flat and a clearance hole ½" diameter in the earthed plate is required to prevent flash-over. Such a condenser will convey at least 10 amps R.F. and will have a capacitance of about 700 pF if mica 0·2 mm. thick is used; this should be adequate insulation for 1,000 volts.

OSCILLATORS

It is not proposed here to discuss in detail the technique of oscillator design because modern radio equipment for communication purposes does not use a valve as an oscillator under optimum conditions for power output.

They are principally employed as a stable source of frequency for transmitters or frequency meters or as a B.F.O. or part of the frequency changer in receivers.

In the above cases a high level of output is often undesirable, but a low degree of harmonic generation is frequently essential and hence the circuit is so designed that the valve swings over a small part only of its characteristic and the output may be calculated as if it were an amplifier providing its own grid excitation.

In general it may be said that the treatment is the same as for a Class C amplifier and the ratings are similar except that a lower efficiency is obtained because the valve has to provide its own grid drive and this drive therefore must be deducted from the power output.

In order to obtain good wave-form and stable operation, and not necessarily the maximum output, it is desirable that the peak stored energy per cycle should be at least twice that fed to the load per cycle. Since this energy is stored in the tank circuit the Q of the circuit allowing for the equivalent resistance of the load should be at least 4π and the inductance and capacitance of the tank circuit are given by:—

$$L = \frac{V_a^2}{8\pi^2 \times W_{out} \times f}$$ henrys, where f is frequency in cycles

$$C = \frac{2 \times W_{out}}{V_a^2 \times f}$$ farads.

CHAPTER 5 DETECTORS

THE attention of those readers who have ambitions towards quality reproduction is particularly drawn to this Chapter. Measurements show that the average detector is a much more fruitful source of distortion than the A.F. amplifier. Typical detectors of all types, as used in normal broadcast or communication receivers, give a harmonic distortion at 100 per cent. modulation of the order of 10 per cent.

It is the object of this Chapter to indicate the methods involved in the choice of operating conditions rather than to compare the merits of various types.

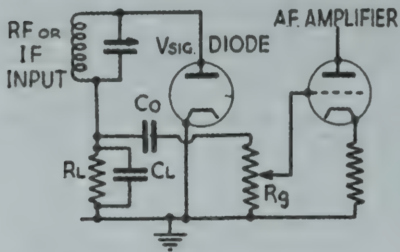


Fig. 22.
Circuit of a conventional diode detector.

Diode Detectors

The circuit of a conventional diode detector is shown in Fig. 22. The signal (V_{sig}) applied to the diode anode from an R.F. or I.F. source, the resistor (R_L) provides the diode load across which the demodulated signal is available; the rectified carrier current also produces a D.C. voltage. Across this load is shunted a condenser C_L to provide an easy path for R.F. but not A.F. currents. The coupling condenser C_0 prevents the D.C. voltage appearing across the A.F. volume control R_g and thus biasing the grid of the A.F. amplifier valve. The time constant of C_L and R_L must be long in relation to the carrier frequency but short compared with the modulation frequency.

In order to obtain a high degree of efficiency it is desirable to make R_L high compared with the internal impedance of the diode (r_a), which in the case of a normal receiving diode varies from about 200 ohms for low impedance television types to about 10,000 ohms for those associated with a triode in a double diode triode.

It will be noticed that as in a voltage amplifier the grid leak of the succeeding valve is in parallel with the diode load from an A.C. point of view but in this case it does more than just lower the effective amplification. It will in fact cause the A.F. output to be distorted above certain depths of modulation because the A.C. load is lower than the D.C. load.

Fig. 23 shows the anode current plotted against anode voltage (I_a/V_a) for a typical high impedance detector diode for various R.M.S. input voltages. D.C. load lines have been drawn for different values of diode load (R_L), and it will be observed that with zero input volts and zero anode volts there

is a small standing current. This is due to contact potential within the valve and represents an artificial D.C. bias of about 1 volt. If, for example, a load of 500,000 ohms is used and a carrier of 15 volts R.M.S. is applied the operating point will be at A with the result that an anode current of 39 microamps will flow and a D.C. bias of 19·5 volts will be set up in the load. If now the carrier is modulated at 100 per cent. the carrier will swing about point A from 30 volts (point B) to zero volts (point C). Hence an A.F. voltage will appear as well as the D.C. voltage, the amplitude of the A.F. voltage being proportional to the lengths AB for the positive peak and AC for the negative. The ratio of these lengths is an indication of the distortion; in this case the positive peak is 19·5 volts and the negative 18·5 volts. The R.M.S. A.F. output at 100 per cent. modulation is 13·5 volts.

It can be shown that the maximum percentage modulation that can be accepted without distortion is equal to:

$$100 \times \frac{\text{negative peak}}{\text{positive peak}} = 100 \times \frac{18 \cdot 5}{19 \cdot 5} \text{ or } = 95 \text{ per cent.}$$

In the above case, only the D.C. diode load has been considered, but in practice the A.C. load includes the succeeding valve grid leak or volume control (R_g). If this is taken as 1 megohm the effective A.C. load is 330,000 ohms assuming the reactance of C_o is negligibly low. The D.C. load settles the operating point so the point A remains unchanged, but the A.C. load line will have a slope equivalent to 330,000 ohms instead of 500,000, and is shown as the line EAF. With 100 per cent. modulation the carrier

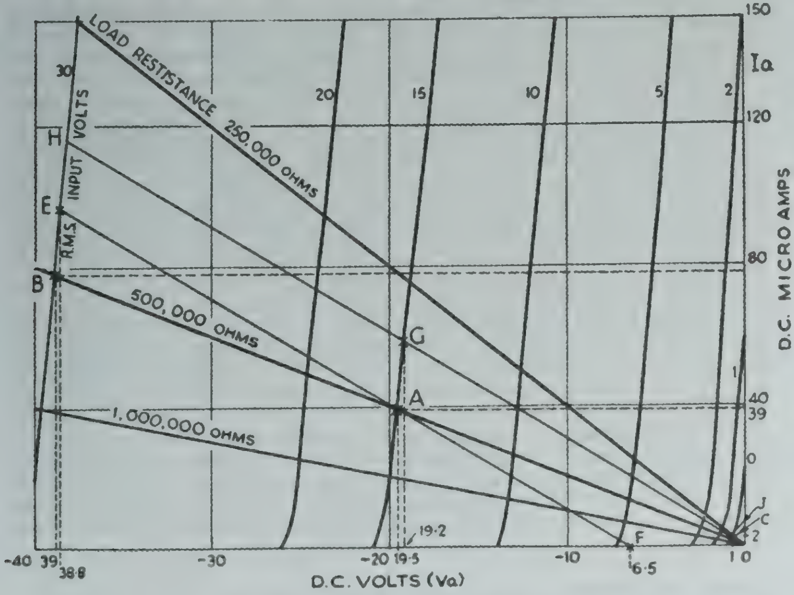


Fig. 23.
Dynamic curves of a typical high impedance detector diode showing various D.C. and A.C. load lines and the effects of a positive bias on the diode for reducing the distortion.

will swing about the point A to 30 volts (point E) and to zero volts (point "F"). It is at once evident that the load line at the point "F" cannot reach the zero input curve; in fact it would only reach an input of about 4 volts. Hence the negative peak of the modulation is trying to take the diode beyond cut-off. The maximum modulation depth which can be accepted without distortion under these conditions, allowing for the A.C. load $= 100 \times \dfrac{19\cdot5 - 6\cdot5}{38\cdot8 - 19\cdot5} = 67\cdot5\%$ and the mean A.F. output at 100 per cent. modulation will be 11·5 volts.

From the above example it is clear that the ratio of the D.C. load to the A. C. load is of considerable importance, and it is essential that the grid leak or volume control have as high a value as is practicable. Unfortunately the maximum resistance that may be used in the grid circuit of a valve is limited by the danger of grid current. Further, if the resistance is a volume control, the input capacity of the valve will cause a variation of the A.F. frequency response with varying volume settings. An alternative would seem to be to reduce the value of the D.C. load keeping the A.C. load as high as practicable. But this results in further difficulties because the lower the D.C. load the more non-linear the D.C. load line becomes—unless a diode is chosen having a lower impedance. Also, as the input impedance of a diode detector is approximately equal to half the value of a D.C. load resistor, a lowering of the D.C. load would heavily load the R.F. or I.F. input, so that it might not then be possible to provide sufficient input from a normal R.F. pentode as an R.F. or I.F. amplifier. If the input is reduced, the contact potential at point C mentioned earlier becomes more important, because the operating point A will move to the right, and since C is fixed the difference in the lengths of the lines BA — AC, will increase at an increasing rate as the signal input is reduced. Actually the greater the input the less the distortion.

There is another matter not yet covered that arises in a practical diode detector circuit. This is that the R.F. or I.F. carrier should not be allowed to reach the A.F. amplifier and a filter circuit must be inserted between the diode load and the grid resistor. This filter usually comprises a series resistance and a parallel condenser combination, but a choke may be used in place of the resistance. If the resistance is too low in value or the choke of too low inductance the filtering will be poor and the following condenser will in effect be in parallel with C_L (see Fig. 22) thus lowering the A.C. load at the higher audio frequencies. On the other hand if the opposite is true the A.C. load will be higher, but the higher audio as well as the radio and intermediate frequencies will be attenuated by the filter. Most of the distortion due to a poor ratio of D.C. to A.C. load may be reduced by the application of a suitable value of positive bias to the diode. The application of +0·3 volts to the diode in the example in Fig. 23 would move the operating point from A to G and the 330,000 ohms A.C. load line would now be HGJ and the maximum modulation acceptable would be again 95 per cent. This small positive bias is critical in value and is only correct for one particular carrier input voltage. The method may be applied to experimental receivers equipped with some method of monitoring the applied input but if applied to receivers incorporating A.V.C. or within a fading area of a broadcast station some automatic method of adjusting the value is essential.

Leaky Grid Detectors

The operation of this type of detector is somewhat complex but is analogous to that of a diode with a direct coupled amplifier. Fig. 24 shows a typical

triode circuit. In such a circuit the grid and cathode may be considered as a diode, the grid being the diode anode, but acting at the same time as the control grid of the triode. R_L, the grid leak, is the diode load and C_L the reservoir condenser. As in the case of the diode detector an input carrier will set up a D.C. voltage across, R_L, which will provide bias for the grid. Both the carrier and the A.F. modulation will be amplified in the triode hence a filter circuit will be required in the anode circuit to prevent the carrier reaching the remainder of the A.F. amplifier.

Considering the grid-cathode as a diode, more severe limitations apply than in a normal diode detector because there is still a contact potential existing between the grid and cathode, but it is not possible to apply nearly as much input as the earlier example because the D.C. voltage developed will bias the grid as an amplifier beyond cut off, with a result that there will be little if any output. It is evident that the ideal input is such that the D.C. voltage developed by the carrier will just bias the grid to the straight portion of its characteristic.

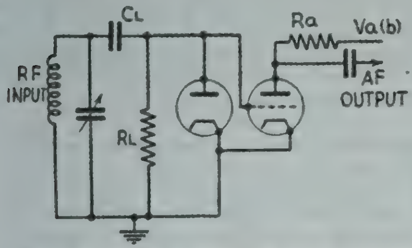


Fig. 24.
A circuit showing the functional operation of a triode as a leaky grid detector.

If curves of the grid voltage—grid current characteristics of the triode are taken (like those of the diode in Fig. 23) and operating conditions are chosen to fit the triode curves as a voltage amplifier (such as those in Fig. 2) it is possible to design the detector.

A difficulty arises, however, because the contact potential of the grid is affected to some extent by the anode current and anode voltage, both of which are varying with the A.F. modulation due to the presence of the anode load R_a. This contact potential variation will have some effect on the shape of the diode curves, and since the input must be fairly small the distortion due to this contact potential will be considerable in the same way as in the earlier example. Fortunately the ratio of D.C. to A.C. load does not enter into the picture, but the time constant of R_L and C_L and the effect of the carrier frequency filter in the anode circuit must still be considered. Too little filtering to the carrier frequency will reduce the efficiency of the amplifier and too much will again attenuate the higher audio frequencies.

The efficiency will be low without an anode by-pass condenser because if the carrier appears across the anode load resistor the " Miller " effect will load the grid with a considerable capacity, which becomes shunted across the grid leak R_L and is in series with C_L. When the anode is by-passed to earth for R.F., the input capacity does not contain a term which is multiplied by the voltage gain.

Because of the small input that could be accepted by conventional detectors of this type, the " power grid " detector was introduced. This is really the same but it employs a triode having a lower amplification factor and a

greater H.T. voltage and in consequence because of its longer grid base it can accept a larger input whilst still operating as a direct coupled amplifier on the linear part of its characteristic. Hence there is less distortion in the "diode" part, but generally less sensitivity because of the lower "amplifier" gain.

Where tetrodes or pentodes are employed, the operation is similar and approximate calculations may be made from a study of the grid to cathode characteristics in the case of a diode and the dynamic I_a/V_a curves in the case of a voltage amplifier.

As is well-known the operation of such detectors is affected considerably by the application of bias to the grid leak, which may be obtained by returning the grid leak to the positive filament lead when battery valves are used. This procedure is analogous to the example of a diode with a positive bias shifting the operating point and the load line to a more advantageous position.

Anode Bend Detectors

This type of detector depends for its action on the curvature in the anode current/grid voltage (I_a/V_g) characteristics. If a valve is biased nearly to cut-off and an input is applied to the grid, the negative half cycle will drive the grid more negative and only a small change in anode current will occur, whereas the positive half cycle will drive the grid positive and cause a considerable increase in anode current.

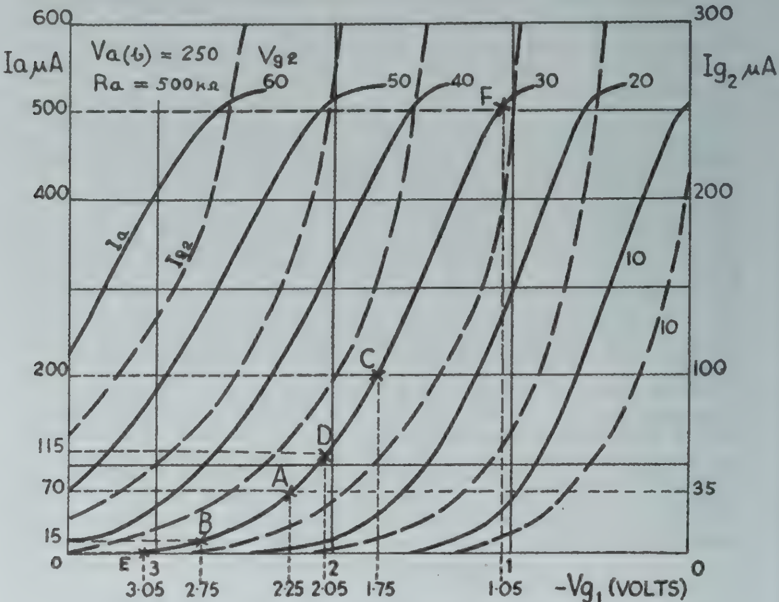


Fig. 25.
Dynamic curves of an R.F. Pentode showing the operating conditions when used as an Anode bend detector.

For an anode bend detector, a valve having a short grid base and as sharp a bottom bend as possible should be chosen. The valve may be coupled to the following stage by any method applicable to voltage amplifiers.

As an example Fig. 25 shows the dynamic I_a/V_g curves of the R.F. Pentode used in the voltage amplifier shown in Fig. 5, but in this case the operating point should be near cut-off. The valve is assumed to be R.C. coupled as before with a 500,000 ohms anode load. A curve should be chosen so that the grid is always more than one volt negative to the cathode, otherwise grid current will flow and cause damping to the R.F. input circuit. Such a curve is that for screen volts 30. A possible operating point is –2·25 volts at A. If an unmodulated carrier of 0·5 volts peak is applied to the grid the operating point will swing from –2.75 volts at B to –1.75 volts at C. Swinging from A to B will decrease the anode current from 70 to 15 microamps (55 microamps) whereas swinging from A to C will increase the current from 70 to 200 microamps (130 microamps). Since the anode current increase is greater than the decrease, rectification will occur and also the mean anode current will increase. The increase will be proportional to the ratio of the areas under the curves for the portion A B and A C. The ratio in this case is 1·65: so that the anode current will increase to 115 microamps and the operating point will effectively shift to D.

If now a 100 per cent. modulated carrier is applied the swing will be 1 volt peak, in other words from –3·05 to –1·05 volts E to F. The mean anode current will further increase in proportion to the areas under this curve and the detector efficiency will also be proportionate to these areas. If automatic bias is employed the resistor in the cathode should be chosen to fit the curves at point A where the anode current is 70 μA. The screen current is 35 μA. so the cathode current is 105 μA., and since the required voltage is –2·25 volts, the resistor will be $\frac{2 \cdot 25}{105} \times 10^6 = 21{,}000$ ohms.

It is quite evident from the appearance of the curve that over D E and D F some considerable distortion will result since the negative half cycle of the modulation will not be completely suppressed, because neither the portion of curve B E nor the portion D F is linear.

A valve having a much sharper cut off is really required but in general such valves do not exist. Measurements show that distortion, of the order of 10 per cent. on 100 per cent. modulation will occur in an anode-bend detector valve under the conditions quoted.

The peak output voltage (for an input of 0·5 volts swing at 100 per cent. modulation) will be somewhat less than half the length D F represented by a change of 115 to 500 μA. in 500,000 ohms $= \frac{1}{2} \times \frac{385}{10^6} \times 500{,}000 = 96$ volts peak or 68 volts R.M.S. This figure is optimistic because due allowance has not been made for the fact that the negative halfwave is not entirely suppressed and this should be deducted from the positive halfwave to obtain the true output. Also, in the same way as for voltage amplifiers, due allowance should be made for the succeeding A.F. valve grid resistor.

Infinite Impedance Detectors

If the fundamental anode bend detector is re-arranged in such a manner that the output load is in the cathode of the valve instead of the anode a variety of " cathode follower " results. Because the input impedance is very

high due to the 100 per cent. negative feed-back, it is referred to as an infinite impedance detector. Fig. 26 shows the essential circuit. The modulated carrier is applied to the grid and the load in the cathode is R_k. This load is by-passed by a condenser C_k, which should have a reactance low to carrier frequency but high to modulation frequency. The anode is also by-passed with a condenser C_a large enough to be a low reactance to both frequencies. The rectified output is available across the resistor R_k and is supplied to the A.F. amplifier via a filter circuit comprising C_o, R_o and C_g and R_g. Such a filter is generally necessary because there is often quite an appreciable carrier voltage remaining across R_k despite the effect of C_k. The resistor R_o may be replaced with advantage by a H.F. choke, if difficulty is experienced with drop in response at the higher audio frequencies.

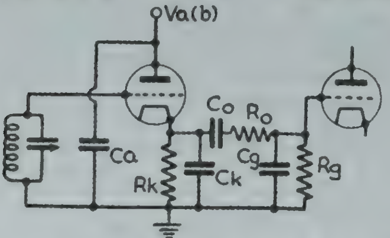


Fig. 26.
The essential circuit of an infinite impedance detector.

The resistor R_k should have a high value compared with the reciprocal of the mutual conductance of the valve. Since for most values $1/g_m$ is of the order of 100 to 1,000 ohms suitable values for R_k are from 10,000–100,000 ohms. The gain of such a detector, like all " Cathode follower " circuits, is less than unity. Neglecting inter-electrode capacities the gain = $\frac{R_k}{1000/g_m + R_k}$ and the input impedance = $c_{ga} + \frac{c_{in}}{\mu}$ where μ is the valve amplification factor at the operating point. Fig. 27 shows the anode current/anode voltage (I_a/V_a) curves for the triode used as a voltage amplifier example in Fig. 2 having a rated amplification factor (μ) of 20, slope of 2·4 mA./V. at V_a = 250 and V_g = –8. The value of $1/g_m$ is approximately 400 ohms so that a cathode load of 25,000 ohms would be quite sufficient.

Using an H.T. line voltage ($V_{a\ (b)}$) of 250 volts the load line will be A B. The fact that the load is in the cathode instead of the anode makes no difference to the load line, but the voltage drop across the cathode resistor is applied as grid bias hence the operating point will have to be such that the anode current multiplied by R_k is equal to the grid voltage and this voltage must lie on the line AB and yet fit the characteristic curves. Such a condition is grid volts –13 where the anode current is 0·52 mA, the drop in R_k being 13 volts and that in the valve (250–13) or 237 volts. The point C is obviously close to the cut-off of the valve and any appreciable input to the grid would result in cut-off being reached on one half cycle in the same way as for an anode bend detector. If, for example, an input of carrier of 3 volts peak is

applied the grid would appear to swing from —10 volts to —16 volts, but in fact it will not do so, because the increase in anode current will change the grid voltage, as the input is not being applied between grid and cathode. For example at –10 volts, the anode current would be 1·6 mA, the drop in R_k 40 volts, and the drop in the valve 210 volts. Clearly, all three cannot happen at one time. The trouble is that a different sort of curve is required, a dynamic one obtained by plotting applied voltage against anode current rather than actual grid voltage against anode current. Such a curve can easily be obtained for any given value of R_k by replotting Fig. 27. For example a grid voltage of zero (point D) gives an anode current of 7·1 mA or an input voltage of +177·5 volts. A grid voltage of –10 (point E) gives 1·6 mA or an input of (40 –10) or 30 volts. A grid voltage of –13 (point C) gives 0·52 mA or input zero and a grid voltage of –16 (point B) gives zero anode current and hence corresponds to an input of –16 volts. A portion of such a replotted curve is shown in Fig. 28a with the equivalent points B C and E marked. Other curves have been drawn for values of R_k of 10,000 and 50,000 ohms. It is clear that the curve is almost perfectly straight indicating low distortion.

Between points C and E, the anode current increases from 0·52 mA to 1·6 mA for a change of 30 volts in input voltage. Hence the change in

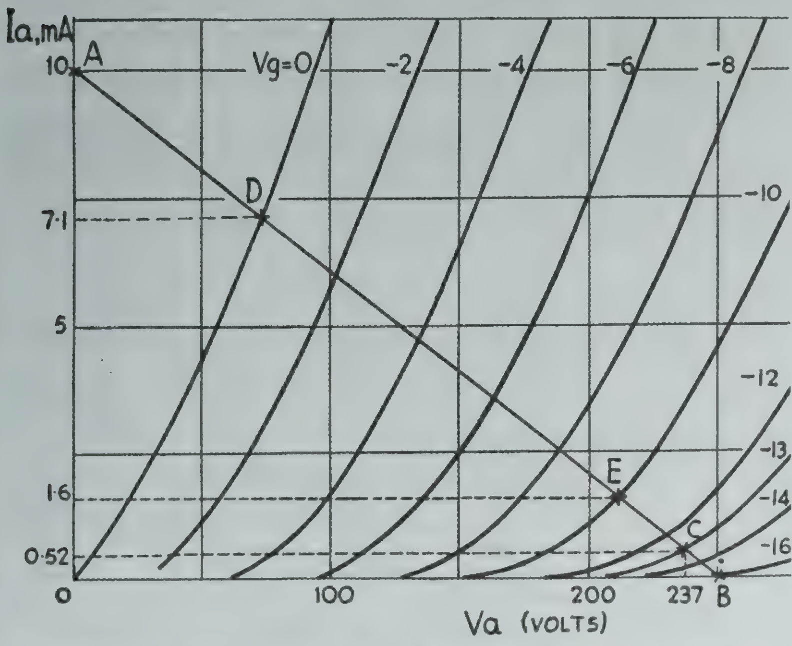
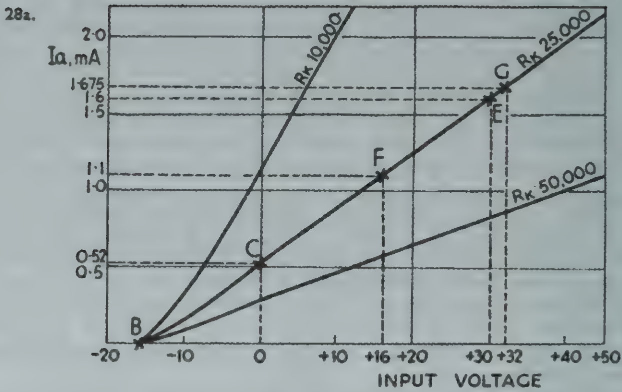


Fig. 27.
The load line as an infinite impedance detector drawn on the characteristic curves of a typical triode.



cathode voltage (or output) is $\frac{1 \cdot 08 \times 25{,}000}{1{,}000}$ or 27 volts. The voltage gain is, therefore, $\frac{27}{30}$ or 0·9. If a carrier of 16 volts peak amplitude is applied the anode current will swing from B to a point F at I_a 1·1 mA and twice the carrier, 32 volts, to point G.

A family of dynamic curves can be taken if various A.C. inputs are applied to the detector and the anode current is plotted against D.C. voltage drop across the cathode resistor (R_k). Such a family of curves is shown in Fig. 28b.

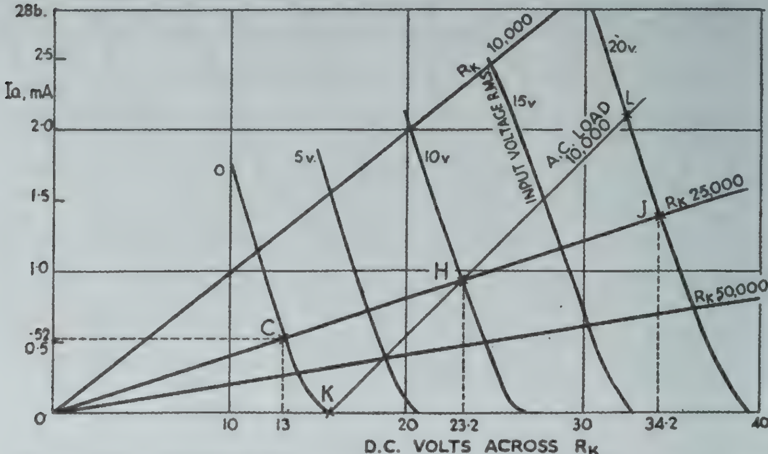


Fig. 28.
(*a*) A dynamic curve for an infinite impedance detector re-plotted from Fig. 27. (*b*) A family of dynamic curves for an infinite impedance detector.

The curve for zero input is the standing anode current with no signal applied. D.C. load lines have been drawn on the curve for loads of 10,000, 25,000 and 50,000 ohms. It will be noted that the anode current with zero volts input on the 25,000 ohm line is 0·52 mA which is point C. These curves may be examined using any suitable sine wave input, such as 50 cycle mains, providing the anode supply is adequately decoupled to 50 cycles and the time constant of C_k R_k is suitable. If R_k is 25,000 ohms, C_k may be $2\mu F$ at 50 cycles.

It is seen that an increase in A.C. input increases the anode current, indicating anode bend rectification. If a carrier of 10 volts R.M.S. is applied, the anode current increases to point H, which is the operating point for this carrier voltage. When this carrier is modulated 100 per cent. the operating point will swing over the portion of the curve from J to C, the distortion being indicated by the difference in the lengths CH and HJ. These lengths represent changes of 10·2 and 11 volts across R_k or a mean of 10·6 volts. This figure is the peak A.F. output with 10 volts R.M.S. input 100 per cent. modulated, so the R.M.S. A.F. output will be 10·6 × 0·707 or 7·5 volts.

So far it has been assumed that the value of R_k represents both the A.C. and the D.C. load for the detector, but this may not be so. For example, in Fig. 26 if the impedance of the filter and grid leak combination C_o, R_o, C_g and R_g was 17,000 ohms to A.F. then the A.C. load line would be 17,000 in parallel with R_k or 10,000. Such a line KHL has been drawn about the point H and this line represents the lowest A.C. load possible without travelling beyond anode current cut-off with a 100 per cent. negative modulation peak so that this detector can accept a ratio of A.C./D.C. load of $\frac{(10,000)}{(25,000)}$ or 0·4 for 100 per cent modulation.

In practice it is unlikely that A.C./D.C. ratio would be low as this figure so that this detector is not normally subject to such difficult requirements as regards the ratio as is the diode. It should be noted, however, that the distortion is increased with a low A.C. load since the ratio of the lengths HL/KH is greater than that of HJ/CH.

In practice a detector using a value of R_k of 25,000 and A.C. load of 250,000 ohms with an input of the order of 10 volts R.M.S. will produce a total harmonic distortion not exceeding 3 per cent. at 100 per cent. modulation. Inputs less than about 5 volts are not advised as the curves converge closer and closer together giving increased distortion.

The value of C_k must not be made too small or the valve will appear as a negative resistance input to the tuned circuit, causing actual instability, or at least such an increase in "Q" of the coil that the side bands will be cut, due to the increased selectivity. A suitable value for C_k (with R_k = 25,000 ohms) is from 250 to 500 p.F. at 1 Mc/s.

CHAPTER 6 FREQUENCY CHANGERS

IN order that a valve may act as a frequency changer it is essential that it shall operate over a non-linear part of its characteristic. This infers that the process is akin to signal detection and any type of valve that will operate as a detector will operate as a frequency changer. From this it does not follow that they will be efficient. The essentials are that an input signal is applied together with a heterodyne voltage of such amplitude that the valve swings over a non-linear part of the characteristic with a result that the sum and difference frequencies appear in the output circuit. The overall voltage gain of a frequency changer or the ratio of I.F. voltage output to R.F. signal voltage input is known as the "conversion gain" and the ratio of the beat frequency component in the output current $(f_1 - f_2)$ or $(f_1 + f_2)$ to the input voltage f_1 is known as the "conversion conductance" (g_c) and is usually stated in microamperes/volt.

The conversion conductance is measured by applying an A.C. voltage to the control electrode and reading the change in D.C. anode current. Thus it is evident that if the valve is operating over a straight portion of the characteristic the change will be zero and no beat frequency product will result.

In order to obtain a more accurate figure for g_c it is usual in the case of multi-grid valves to make a known change in the voltage of the control electrode in either direction whilst injecting a suitable value of voltage on the oscillator electrode to give the working value of heterodyne voltage. The D.C. anode current will then change both up and down compared with the mean value and the value of g_c is half the total change in anode current.

In general it is desirable that the input signal voltage shall be small compared with the heterodyne voltage, which is usually the case because the frequency changer valve is essentially one of the early stages in the receiver circuit.

Figures for conversion conductance may be obtained by using 50 cycle mains for the signal and heterodyne voltage. Care must be taken that all voltage supplies are decoupled with large condensers of low reactance at 50 cycles, and other condensers in grid or cathode circuits are increased in value proportionately to the frequency. For example 100 pF used at 1 Mc/s. would be equivalent to 2 μF at 50 cycles. A suitable signal input is 1 volt and the small changes in anode current may be read on a low range meter, biased back, such as used in many valve voltmeter circuits.

The voltage gain of a frequency changer is approximately the conversion conductance multiplied by the dynamic resistance of the I.F. transformer primary. Due allowance must be made in calculating the "Q" and the selectivity of the I.F. transformer for the anode-cathode impedance of the valve in shunt with the primary. If it is not possible to measure the conversion conductance the gain of the stage may be calculated as if it were an I.F. amplifier operating at the same point, and as a rough guide the gain as a frequency changer will be from 0·4 to 0·6 times the I.F. gain.

The types of frequency changer circuit are legion but descriptions of the basic types follow.

Diode Frequency Changers

A diode may be used as a frequency changer, a typical circuit being shown in Fig. 29. The R.F. input is applied to the anode, the heterodyne voltage to

the cathode and the I.F. output is taken from the cathode. The resistor R_k and condenser C_k provide the diode load. The diode load is used to fix the operating point, and to prevent the diode from acting as a short circuit to the signal, the oscillator and the I.F. amplifier. The impedance presented to all three is approximately equal to half the value of R_k. The value of the reactance of C_k should be low compared with the frequency of either.

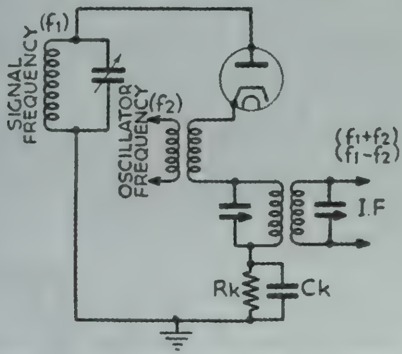


Fig. 29. Typical circuit of a diode valve used as a frequency changer.

In order to determine the operating conditions reference should be made to the method described for diode detectors.

A load line is drawn as in Fig. 23, and the heterodyne voltage should be reasonably large compared with that due to contact potential. Usually about 3 volts or upwards is adequate unless the heterodyne frequency is a harmonic of the local oscillator in which case it can be larger. The slope of the load line in terms of anode current (in microamps) per volt of input signal is the conversion conductance g_c. For example, for the 500,000 ohm line in Fig. 23, an input of 15 volts produces 39 microamps. Therefore the slope is 2·6 microamps/volt in the D.C. load. The A.C. load is the impedance of the primary of the I.F. transformer shown in Fig. 29 and will be drawn at the operating point in same way as EAF in Fig. 23, assuming the impedances of the signal input circuit and the oscillator injection circuit are low at I.F. This type of frequency changer is generally employed at U.H.F., where difficulties arise with more complicated types. It has the advantage of very low noise and an almost complete indifference to the value of heterodyne voltage, providing this voltage is above the minimum figure. It will also operate satisfactorily on harmonics of the local oscillator. For U.H.F. operation a low impedance low capacity diode of the television type is generally used and the values of R_k should be about 100,000 ohms and C_k 0·01 μF.

Triode Frequency Changers

These are usually of the anode bend type, the signal being applied to the grid, the oscillator injected in series with the cathode, and the I.F. transformer being connected in the anode. The anode should be by-passed to ground with a condenser of low reactance at signal, and oscillator frequencies. The capacities used for this purpose will comprise part of the I.F. tuning.

Fig. 30 shows a typical circuit, the resistor R_k and condenser C_k are now the autobias circuit, the reactance of C_k being low to all three frequencies.

A value of heterodyne voltage should be used such that the grid swings from beyond cut off to nearly the point of start of grid current such as from E to F in Fig. 25, the slope between C and F being 300 microamps for 0·3 volts or a g_c of 1,000 μa/V. This curve of Fig. 25 is only taken as an example and it should be noted that this was plotted with a resistor of 500,000 ohms in the anode whereas in practice a load line must be drawn on the normal I_a/V_a curves corresponding to the impedance of the I.F. transformer primary as the load in order to determine the I.F. output voltage and conversion conductance.

Triodes are used in the above manner for short wave receivers where a very good signal to noise ratio is essential and where there is considerable difference in frequency between the signal frequency and the I.F. Care should be taken to match the anode impedance to the I.F. transformer.

The value of g_c will generally be approximately one quarter of the value of the g_m used as an amplifier.

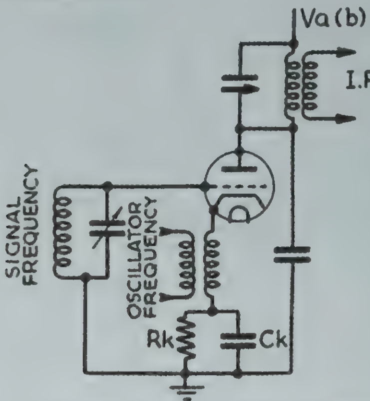


Fig. 30. A triode valve used as a frequency changer.

Pentode Frequency Changers

A pentode valve may be employed as a frequency changer in a number of different ways. In general the signal is always applied to the control grid g_1 and the I.F. output is taken from the anode circuit, but the heterodyne voltage may be injected in the control grid or cathode circuit, when the valve functions in the same way as an anode bend detector. The same considerations apply as for the triode frequency changer, except that due to the much smaller control grid-anode capacity, interaction between the signal and I.F. circuits if they are more nearly alike in frequency, is less troublesome. In addition the heterodyne voltage may be injected at the screen g_2 or the suppressor grid g_3.

In the former case Fig. 31 shows the curves of a typical pentode plotting anode current and screen current against screen voltage for a constant g_1

voltage, and mutual conductance against screen voltage ; from these it is evident that if the screen voltage is varied the anode current and mutual conductance will vary also. Thus if the screen voltage is fixed at 60 volts as at point A, and a heterodyne peak voltage of 20 volts is applied, the screen voltage will vary from 40 to 80 volts and the anode current from 0 to 0·98 mA at C. Since the increase in anode current will be greater than the decrease, the valve will act as a frequency changer. The mean anode current will as a result increase proportionately to the areas under the curve to a point about B, 0·67 mA.

As at this point the slope is also 0·67 mA/V the conversion conductance will be about half this figure. The actual value may be measured by applying a 1 volt change to the grid in either direction and noting the change in anode current whilst the screen is maintained at 60 volts D.C. with a 20 volts A.C. applied in series with it. The disadvantage of this method is the large heterodyne voltage required for good conversion.

Fig. 32 shows the curves of an R.F. pentode plotting the anode current against suppressor grid voltage (I_a/V_{g3}) with constant screen and control

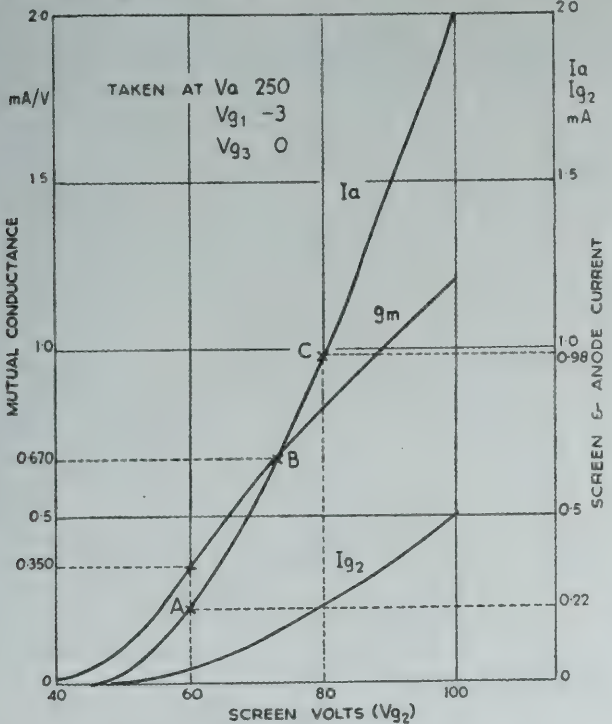


Fig. 31.

Characteristic curves of a typical R.F. pentode plotting anode current against screen voltage and showing their use for determining the performance of a pentode frequency changer with screen grid injection.

grid voltage. There are two major bends in the characteristic and either may be used. The suppressor may be biased to a point such as A and a heterodyne voltage of the order of 25 volts applied swinging the anode current from B to C, or a large bias may be used (point D) when the current will swing from E to F. If the latter point is used care must be taken that screen current is not so excessive as to cause the safe screen dissipation to be exceeded. The determination of the corresponding drop or rise in anode current is performed in the normal manner and the conversion conductance may be determined in the way previously described. This form of injection, like screen injection, requires a large heterodyne voltage and if an appreciable amount of this voltage is allowed to be picked up by the control grid circuit either by valve capacities or layout, etc., this voltage on the control grid will be amplified by the gain of the valve considered as a triode (the screen acting as anode) and may be comparable with that being intentionally applied to the suppressor. Since it will be of opposite phase demodulation will take place and the conversion will be poor. This effect is particularly troublesome at high frequencies.

Multi-grid Frequency Changers

These types of valve are specifically designed to act as frequency changers. They are designed to have a reasonably high conversion conductance, with as low a value of heterodyne voltage as is practicable and a high anode

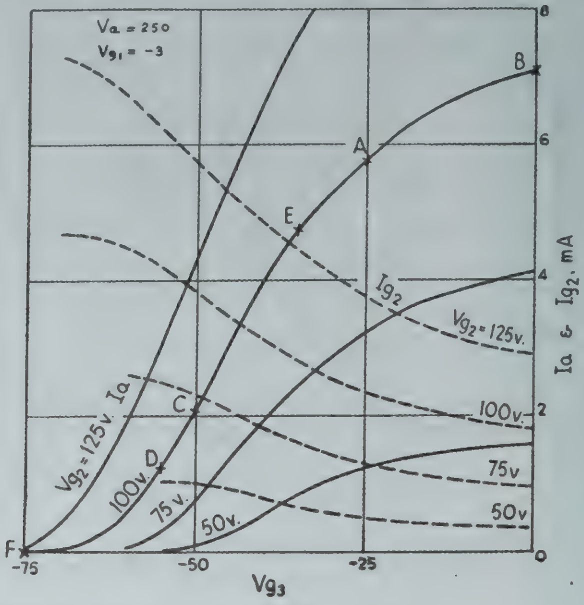


Fig. 32.
Characteristic curves of an R.F. pentode plotting anode current against suppressor grid voltage and showing their use for determining the performance of a pentode frequency changer with suppressor grid injection.

impedance so as not to damp the I.F. transformer unduly. The signal frequency may be taken close to the I.F. without difficulty and A.V.C. may be employed satisfactorily. The only disadvantages are their limited performance at high frequencies and the fact that the signal to noise ratio is poor. For a further discussion on the latter point the reader is referred to a later section of this book.

Some types are designed for use with a separate oscillator to provide the heterodyne voltage whilst others have a self-contained triode to provide such voltage. Certain types of heptode are arranged to oscillate as an electron coupled oscillator with part of the oscillator coil in the cathode circuit. The essential point is that the oscillator must be arranged to provide the necessary heterodyne voltage and as far as practicable this voltage should remain constant over any given frequency band.

There are two types of essential characteristics published by the valve manufacturers, and the first is shown in Fig. 33 for a typical triode hexode valve. This plots conversion conductance, anode impedance, and cathode current against heterodyne voltage at the operating control grid bias. The "X–axis" may be shown in various ways. The scale may be as shown, which

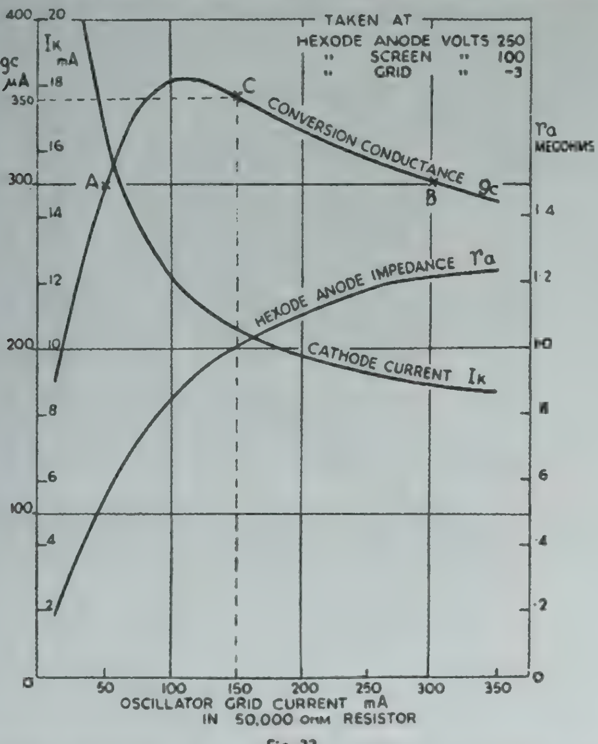


Fig. 33.
Typical published characteristics of a triode-hexode frequency changer showing effects of varying the heterodyne voltage.

is the D.C. grid current in the recommended value of grid leak, which is a simple measure of the heterodyne voltage, or it may be directly as values of heterodyne peak voltage or as heterodyne R.M.S. voltage, but in any case such curves show the relation between the g_c, r_a or I_k and heterodyne voltage.

It will be noticed that for small values of oscillator grid voltage the value of g_c is falling very rapidly, as is the value of r_a, so that the gain in a receiver will fall off rapidly with low values of heterodyne voltage. It is therefore desirable that the coupling of the oscillator coils is so adjusted that the voltage never falls below a value corresponding to point A on the curve. Too much voltage causes a less rapid fall in g_c and this fall is to some extent offset by a rise in r_a giving a rising I.F. gain due to reduced loading on the I.F. transformer. Nevertheless it is undesirable for the voltage to exceed a point around B because excessive voltage will tend to generate harmonics making a receiver prone to spurious whistles. If the valve is operating at point C with a grid current of 150 μA the D.C. grid voltage will be 7·5 volts,

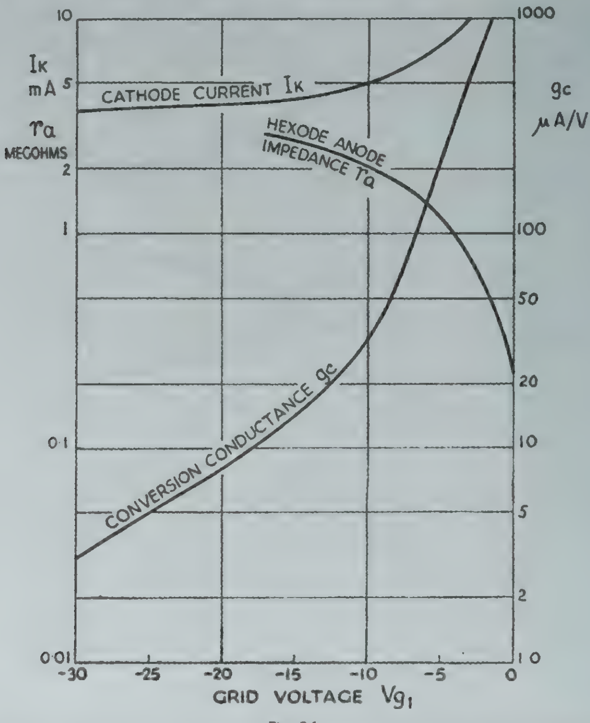
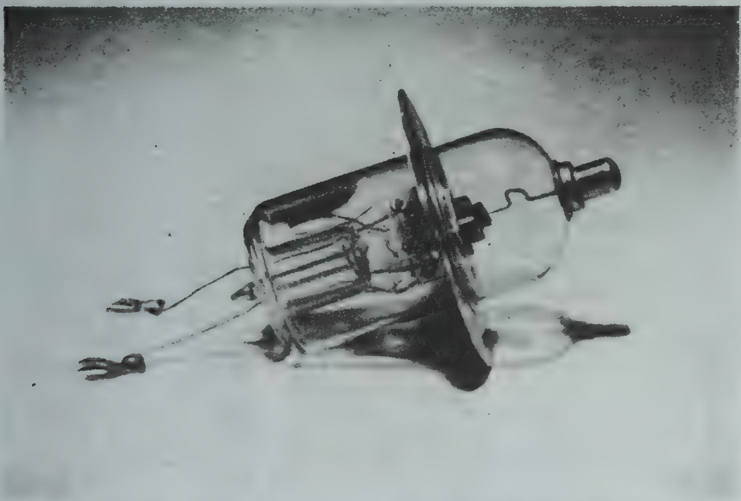


Fig. 34.
Characteristics of a triode hexode frequency changer showing the effects of varying the control grid voltage.

the heterodyne peak voltage will be 1·2 times the D.C. value or 9 volts, and the anode impedance will be 1 megohm. This value is relevant to the design of the I.F. transformer and will determine its step-up. The cathode current is 10·5 mA and hence for a working control grid bias of −3 volts the bias resistance should be 285 ohms. The conversion conductance is 350 μA/V so that if the I.F. coil has a dynamic resistance of say 100,000 ohms the conversion gain will be 35.

The second curve plots g_c, r_a and I_k against control grid voltage for a fixed value of heterodyne voltage, that of a typical valve being shown in Fig. 34. It will be seen that increasing the bias decreases the g_c at a fairly linear rate at the same time raising the value of r_a. This curve enables the gain with A.V.C. voltage to be calculated at any point and also some idea to be obtained as to the maximum signal with A.V.C. that can be applied without cross-modulation.

Essentially similar curves are published for all types of multi-grid frequency changers.



S.T. & C. type 3A/146J grounded grid triode. The top cap is the anode, the grid is part of the copper disc brought out through the glass, the edge of this disc is clamped to a metal chassis or screen. The flexible leads are the heater and cathode connections. When mounted through a screen the anode to cathode capacity is 0.035 p.F.

CHAPTER 7 POWER RECTIFIERS

THE following additional symbols are used in this Chapter:—

V_{rms}	R.M.S. voltage of transformer half secondary.
v_{pk}	Peak voltage of transformer half secondary.
I	Load current in amperes.
f	Supply mains frequency in cycles per second.
C	Condenser capacity in farads.
L	Choke inductance in henries.
N	Ripple factor.

Power Rectifier Circuits

There are three important circuits for the supply of direct current for radio and allied purposes from a single phase alternating current source. In order of popularity they are: Bi-phase half wave, half-wave and full-wave. The basic circuits are given in Fig. 35.

Bi-phase Half-wave

The circuit is arranged as in Fig. 35a. The R.M.S. voltage of the combined transformer secondary voltages is approximately twice the unidirectional output voltage and the current loading is such that the winding is not used very efficiently. Each half secondary only delivers current on alternative half waves, but the heating effect instead of being reduced to half the continuous rating is only reduced to $\frac{1}{\sqrt{2}}$ or 0·707 times the continuous rating. Therefore, the transformer must be designed for a continuous rating at least 40 per cent. greater than the required direct current output. The ripple voltage

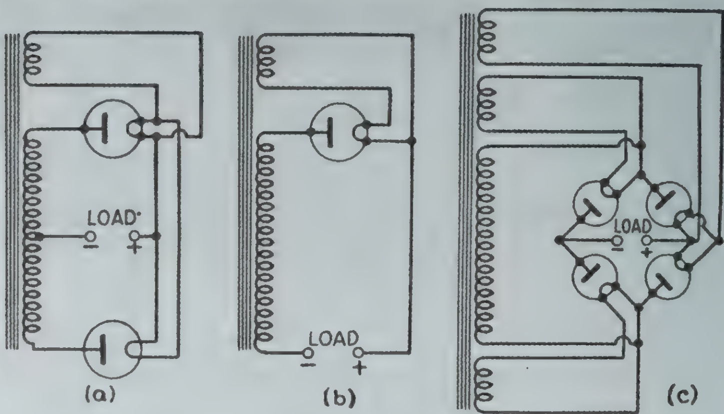


Fig. 35.

Rectifier circuits for single phase A.C. supply. (*a*) Bi-phase half wave. (*b*) Half wave. (*c*) Full wave.

will be at twice the mains frequency and a single filament winding only is required. The peak inverse voltage across the rectifier valve that is non-conducting is approximately equal to the peak voltage of the whole secondary, or three times the direct voltage output. This is the factor that mainly limits the voltage that any rectifier can handle.

Half-wave Rectification

Fig. 35b illustrates this circuit. The R.M.S. voltage is about equal to the output voltage and no transformer secondary taps are needed, but the winding remains inefficiently used as in the previous circuit. Since current flows in only one direction there is also appreciable D.C. magnetisation of the transformer core which further reduces the efficiency. Ripple frequency is the same as the supply frequency which makes smoothing more difficult, but the peak inverse voltage is only half that of the previous case. This circuit is not much used except for very light current applications or for direct rectification of the mains in transformerless power supplies. In the latter case indirectly heated rectifiers are used which are capable of withstanding considerable potentials between heater and cathode, and the heater supply is also obtained directly from the mains through a suitable dropping resistor.

Full-wave Rectification or Bridge Circuit

As in the half wave circuit, the R.M.S. voltage need only be as high as the direct output voltage but the secondary is used to full advantage, current flowing during both half cycles of the alternating current cycle. The rectifier inverse peak is about the same as the peak value of the secondary voltage and again a tapped secondary winding is avoided. Four rectifiers and three filament windings are essential. In the bi-phase circuit it is quite usual to have the two rectifying valves mounted in one bulb using a common filament system, and such a valve is frequently erroneously called a full-wave rectifier. Likewise it is possible to combine two of the four valves used in the full-wave circuit, but the other two must be separate. The output current is only the same as the bi-phase circuit for a given rectifier rating, but since the reverse peak is reduced to half, double the voltage can be obtained without exceeding the valve rating. Ripple voltage is at twice the mains supply frequency.

Smoothing Circuits

The remarks on ripple apply to the case where the output is applied directly to a virtually non-reactive load without the inclusion of any smoothing circuits. Since it is always necessary to reduce the ripple voltage for radio work some form of smoothing circuit will have to be included. The type of smoothing circuit used has a marked effect upon the operating conditions of the rectifier valves and two alternative circuits will be examined in some detail with reference to the bi-phase half wave circuit although the conclusions reached are broadly applicable to both the other circuits.

Condenser Input Filter

This is by far the most common arrangement and is shown in Fig 36a, while Fig. 36b, illustrates some of the wave forms that exist in various parts of the circuit. The line marked V_{a_1} is the voltage applied to the anode of valve $V1$ from one half of the secondary while V_{a_2} is the voltage applied to the anode of valve $V2$ from the other half of the secondary, only the potentials in

the positive direction being shown. The voltage wave of the cathodes and of the positive condenser plate is marked V_K. Starting from point A the condenser discharges into the load and its potential falls along the line AB. At B the anode-cathode potential of rectifier $V2$ is zero and a fraction of a second later the anode is positive to the cathode, and current flows through the valve, charging the condenser and raising its potential along the line BCD. After the point v_{pk}, the anode potential is falling and the cathode potential still rising so that at point D the valve becomes non-conductive and the discharge of the condenser recommences. It is clear from this that current can only flow through the valve for a very small part of each cycle and the peak current through the valve must be several times greater than the steady direct current output. It is necessary to ensure that this peak current will not exceed a safe value for the valve. It is impracticable to get an exact mathematical solution for the magnitude of the peak current or of the ripple voltage, but by making two rather arbitrary assumptions an approximate solution, near enough for normal conditions, can be obtained. The first assumption is that the resistances of the valve and transformer are negligible. The waveforms are then modified as shown in Fig. 37. The lines B v_{pk} D and BCD of Fig. 36 diverge because of the potential difference necessary to overcome the resistance of the valve and transformer, when

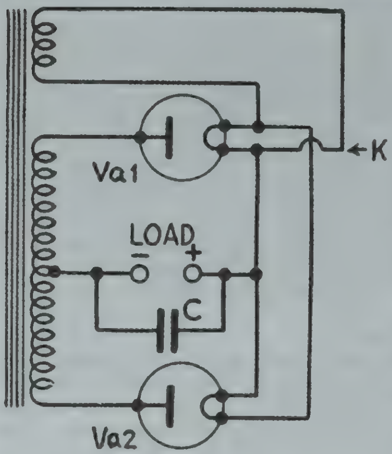


Fig. 36a.
Basic Circuit of full wave rectifier with condenser input filter.

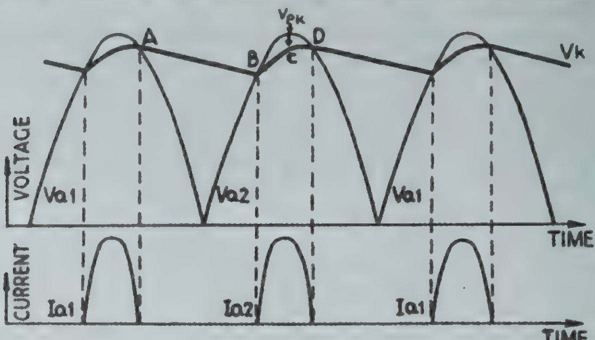


Fig. 36b.
Full wave rectifier with condenser input filter. Current and voltage wave forms.

current flows. In the resistanceless case they coincide. The second assumption is that the charging current, during the period of conductivity, is constant. Introduction of series anode resistances external to both the valve and the transformer is quite a normal method of reducing the magnitude of the peak current in cases where it is found to be too high, so that any calculation based upon an assumption of no resistance is likely to give too high a value for the charging current. On the other hand, the current during the conductivity period is not in practice constant but has a saw tooth form, so that the constant current assumption would give too low a value. Thus the errors tend to cancel, making the simpler solution sufficiently near to be of considerable help in design work.

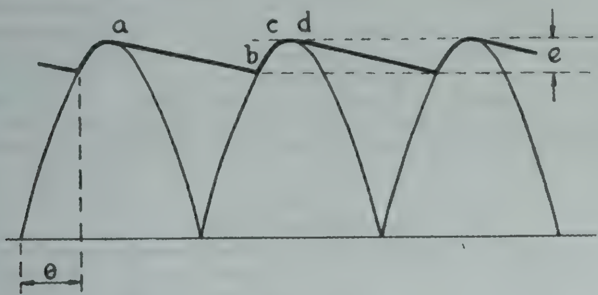


Fig. 37.

Voltage wave form of condenser input filter when valve and transformer are assumed resistanceless.

The formulæ are:—

$$\text{Sin } \theta = 1 - K - \frac{2K\theta}{\pi}$$

Where θ is the angular displacement at the onset of conductivity and $K = \frac{I}{4fC \quad V_{pk}}$

Ripple voltage, peak to peak $= e = V_{pk}\,(1 - \text{Sin } \theta)$

Output voltage nearly equals $V_{pk} - \frac{e}{2}$

Peak current $= I \times \dfrac{\pi}{\frac{\pi}{2} - \theta}$

Ripple factor $N = \dfrac{\text{R.M.S. value of alternating Ripple voltage}}{\text{direct output voltage}}$

$$= \frac{e}{\sqrt{2}\,(2v_{pk} - e)}$$

Illustrative examples have been worked out for three values of current, with a constant condenser size, and for three values of condenser with constant load current. The transformer peak voltage V_{pk} is taken as 500 volts, equivalent to 355 volts R.M.S. The results are shown in Tables 1 and 2.

TABLE 1.

Input Condenser = 8 μf.

Load mA	K	θ	Ripple voltage peak to peak	Output voltage	Rectifier peak current mA
50	0·0625	63°	55	478	335
100	0·125	54°	45	452	500
150	0·1875	46°	140	430	620

Examining these tables, Table 1 shows how the increasing load increases the period of conductivity (θ decreases) and partly offsets the increase in valve peak current that would otherwise be expected. At the same time the ripple is increased and the output voltage reduced. The reduced output voltage is an inherent feature of the circuit which, together with the high peak currents, limit the range of usefulness of the circuit to relatively low powers.

TABLE 2.

Load Current = 100 *mA*.

Input Condenser	K	θ	Ripple voltage peak to peak	Output voltage	Rectifier peak current mA
2 μf	0·5	22°	310	345	265
8 μf	0·125	54°	95	452	500
32 μf	0·03125	71°	30	485	950

Table 2 shows that larger condenser sizes reduce ripple and give a useful increase in output voltage, but put a severe strain on the valve. Makers often specify the maximum condenser size permitted for a given valve, usually about 16 μF. High peak currents are particularly damaging to mercury or gas-filled rectifiers and the condenser circuit is not suitable for use with them. The peaky nature of the current wave further raises the heating effect of the current in the transformer secondary, so that instead of the secondary rating having to be 1·4 times the output wattage a figure of 2·0 or even higher may be needed.

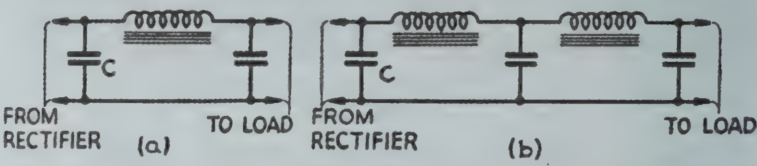


Fig. 38.

Additional smoothing circuit added to condenser input filter. "C" is the rectifier input condenser. (*a*) One section. (*b*) Two section.

Further reduction in ripple voltage over that obtained with a single condenser is usually needed and this can be achieved by adding an extra inductance-capacity section or sections as shown in Figs. 38a and 38b. Each section will reduce the ripple voltage to $\frac{1}{158f^2 LC - 1}$ of its previous value.

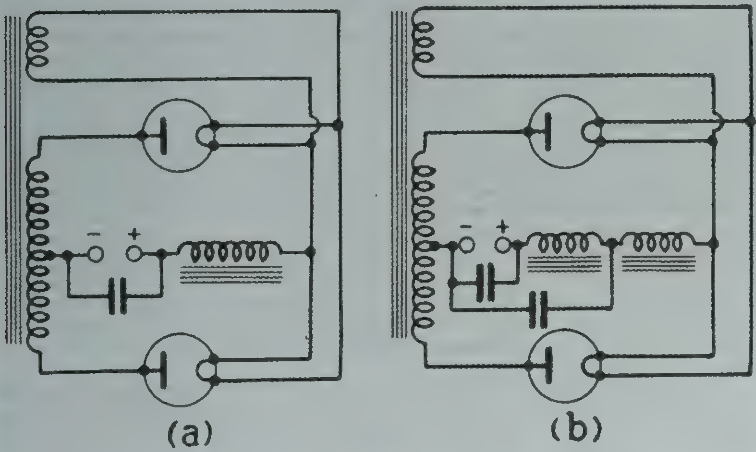


Fig. 39.
Choke input filters. (*a*) Single section. (*b*) Double section.

Choke Input Filter

It is equally possible to smooth the supply by using an inductance followed by a capacitance as in Fig. 39a, more sections being added if necessary to reduce the ripple as in Fig. 39b. Since it is only the first section that has any influence upon the valve operation subsequent sections are ignored in the following paragraphs.

Owing to the reactance of the choke, the effect of the condenser on the direct current performance is practically nil, unless the direct current falls below a certain value. As soon as the direct current becomes less than the peak value of the alternating current flowing through the choke, pulse charging of the condenser commences, the direct voltage output rising steeply on light loads to the same value as in the condenser input case. Provided that this condition is avoided however the rectifier current is continuous and the heavy peak currents, characteristic of the condenser circuit, are avoided. In a normally designed rectifier circuit the peak current should be only about twice the direct current output. The output voltage, ignoring losses in the resistance of the choke, valves and transformer, will be the average voltage of the transformer half secondary, that is the R.M.S. voltage multiplied by 0·9, except in the "light load" region. A comparison of the voltage performance of this type of filter with the condenser circuit is shown in Fig. 40.

The ripple factor N is given by the formula

$$N = \frac{0 \cdot 471}{158 f^2 LC - 1}$$ and the minimum direct current load, to avoid the rising part of the characteristic curve is given by

$$I_{min} = \frac{V_{rms}}{10\,Lf}(0 \cdot 471 - N)$$

Usually N will be small enough to be neglected in this equation, but in the rare occasions when more exact information is required, the formula must be combined with the ripple factor equation and the pair solved as simultaneous equations.

Since it is only the ratio of direct to alternating currents in the choke which is important for good regulation, no harm results if the alternating current increases with the direct current. This is fortunate since by making the choke without an air-gap in the core it is possible to make a reasonably compact component that will have a high inductance at low values of direct current. At higher values of direct current, the core will saturate and the inductance will fall, increasing the magnitude of the alternating current at a time when the regulation will not be destroyed. This type of choke is called a "swinging choke." Whatever the inductance of the choke, however, it will inevitably have a large alternating potential across it and unless it is a well made article with substantial clamps it will emit a most irritating buzz due to lamination vibration.

Each additional choke-capacity filter section, if used, will reduce the ripple voltage by the same amount as given for the subsequent sections in the condenser case, the useful output being reduced only by that lost in the ohmic resistance of the choke.

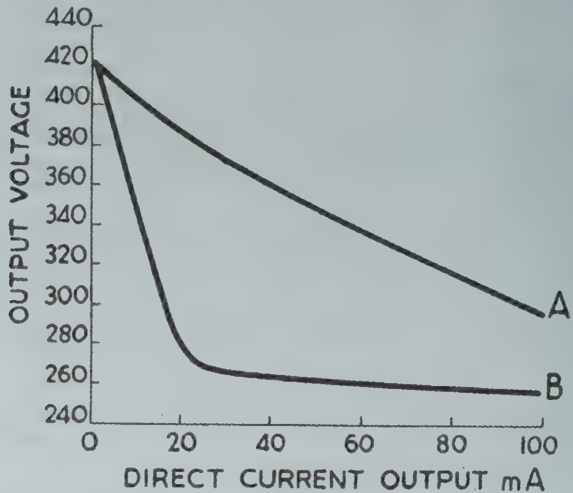


Fig. 40.
Typical regulation curves for rectifiers. (a) Condenser input filter. (b) Choke input filter.

CHAPTER 8 SPECIAL APPLICATIONS

Cathode Followers

IF a valve is operated with the output load in the cathode instead of the anode, this load is common to both grid and anode circuits and the arrangement is called a cathode follower.

Referring to Fig. 41 the input voltage V_{sig} is applied between grid and earth, and the output load R_k is connected between the cathode and earth, across which the output voltage V_{out} is obtained. The anode voltage $V_{a(b)}$ is supplied from a source of low impedance to the signal voltage V_{sig}. If at any instant V_{sig} is made more positive the grid potential will be more positive, the anode current will increase hence the cathode current will increase and the increased voltage drop across R_k will cause V_{out} to be in a positive direction also. From the above it is seen that the output voltage is of the same direction as that of the applied grid voltage, *i.e.* it " follows " the input voltage.

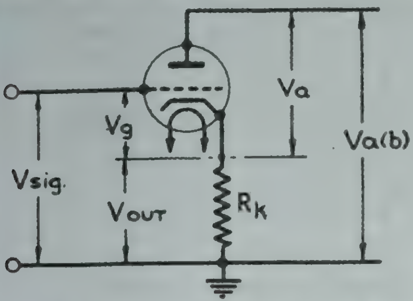


Fig. 41.
Basic circuit of a cathode follower.

The amplification of a *valve* is concerned with the applied voltage between grid and cathode, and this voltage V_g is not V_{sig} but $V_{sig} - V_{out}$. The higher the amplification of the valve the nearer V_{out} approaches in value V_{sig}.

The voltage gain of the stage is $\dfrac{V_{out}}{V_{sig}} = \dfrac{\mu \; R_k}{r_a + R_k(1 + \mu)}$

It is evident that if μ is large, and R_k is large compared with the anode impedance (r_a) then the gain is nearly unity.

Since the load is in the cathode lead both tetrodes and pentodes behave as triodes and the value of μ and r_a are those quoted for the valves connected as a triode.

With the load in the cathode the output impedance is no longer r_a but is reduced by a factor of $(\mu + 1)$ so that the effective output impedance

$$= \frac{r_a}{\mu + 1}$$

If μ is large, the expression approximates to $\dfrac{R_a}{\mu}$ or $1/g_m$. In the same

way the input capacity is reduced and the input capacity of the cathode follower is

$$c_{ga} + \frac{c_{in}}{\mu}$$

Since the cathode follower possesses a low output impedance, a high input impedance and low distortion due to the high degree of negative feed-back with a gain around unity, it behaves as an impedance matching transformer having very little loss over a wide frequency range.

The load in the cathode may be a resistor, a tuned circuit or a loaded transformer, but if a tuned circuit or transformer is employed, and if the impedance of either falls when off resonance, then the circuit will cease to act as a cathode follower.

In order to determine the operating conditions of a valve in a circuit such as Fig. 41, the I_a/V_a curves of the valve connected as triode are required. Such a set of curves is shown in Fig. 42 relating to a tetrode valve connected as a triode. Taking the value of R_k as 2,500 ohms and $V_{a\,(b)}$ as 500 volts,

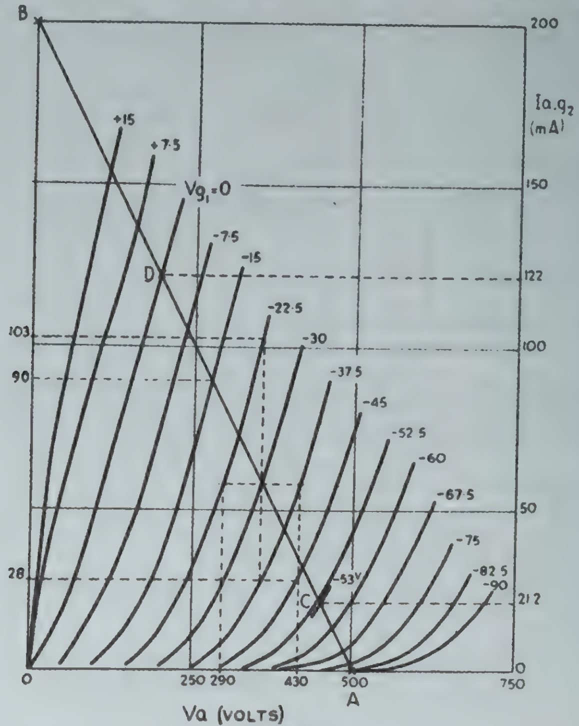


Fig. 42.
Characteristic curves of a typical beam tetrode output valve connected as a triode showing the load line when used as cathode follower.

the load line AB must be drawn from the $I_a = 0$, $V_a = 500$ point with a slope of 2,500 ohms. The curves are for values of V_{g_1}, i.e. the input signal V_{sig} less the output signal V_{out}. The operating point in the absence of any external grid bias will be such that $I_a \times R_k = V_g$ and this point must lie on the line AB and yet fit the characteristic curves. Only one point C satisfies this condition and that is $I_a = 21 \cdot 2$ mA and $V_g = -53$ volts. It is evident that this point is so situated on the load line that a very small signal can be accepted in a negative direction, hence some positive bias is required to bring the operating point nearer the centre of the line. At this point it would be convenient to redraw the curves as a dynamic curve plotting instantaneous values of input signal against anode current. Since the grid must not run into grid current or the input will no longer be high impedance, the grid can only operate down to zero bias point D, and the other extreme being cut-off point at approximately A. A dynamic curve is quickly plotted since at D, I_a is 122 mA, so that V_{sig} is $+305 - 0$ volts; at C, I_a is $21 \cdot 2$ mA so that V_{sig} is $53 - 53$ or zero volts, and at A, I_a is zero so that V_{sig} is $0 - 82 \cdot 5$ or $-82 \cdot 5$ volts. Similarly, at any intermediate point such as $V_g = -15$, $I_a = 90$ mA, V_{sig} is $+225 - 15$ or $+210$ volts. Such a dynamic curve is shown in Fig.

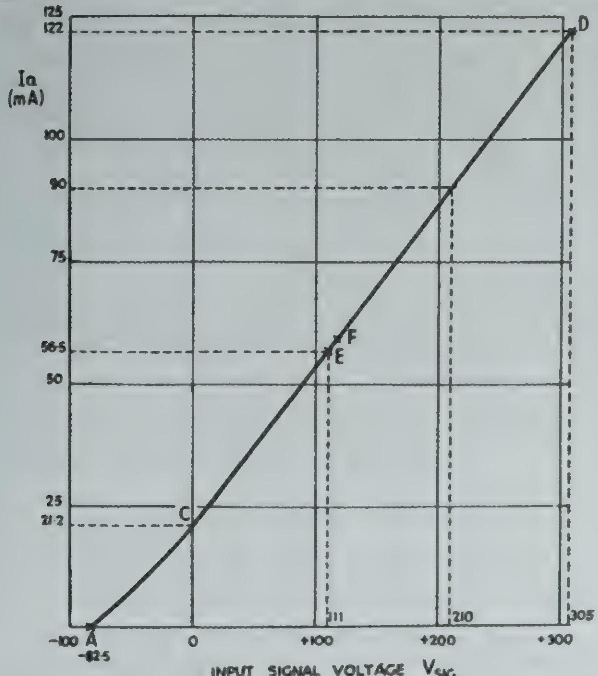


Fig. 43.
A dynamic characteristic showing the operating load line replotted from Fig. 42.

43 with the corresponding points marked. The curve is very nearly a straight line indicating low distortion. The range of input signal is from $-82\cdot5$ to 305 volts, giving a peak to peak value of $387\cdot5$ volts. Hence the maximum input amplitude is half this value or say 194 volts. This signal should be such that the input can swing equally either way, which is $305-194$ or $= 111$ volts point E. At E, I_a is $56\cdot5$ mA so the cathode voltage is 141 volts and the effective grid bias is $111 - 141$ or -30 volts. The operating point E is fixed by applying from some external source $+ 111$ volts to the grid usually from a potentiometer across the H.T. supply via the grid leak or earthy end of the input circuit. The grid leak may be of quite high resistance since the valve at no time draws any grid current.

The gain $= \dfrac{V_{out}}{V_{sig}}$ so that for example between points E and D the I_a change is $122 - 56\cdot5$ or $65\cdot5$ mA, therefore, $V_{out} = \dfrac{65\cdot5 \times 2{,}500}{1{,}000}$ or 164 volts for V_{sig} of $305 - 111$, or 194 volts, hence gain $= \dfrac{164}{194}$ or $0\cdot85$. This figure may be calculated another way because at $V_g = -30$ on the curves (Fig. 42) a change of plus or minus $7\cdot5$ volts in V_g is equivalent to change of 140 volts in V_a so the value of μ is $\dfrac{140}{15}$ or $9\cdot35$. The value of g_m at the same point is a change of I_a of $103-28$, or 75 mA for 15 volts change in V_g, giving a g_m of 5 mA/V. The anode impedance is, therefore, 1,870 ohms. Applying these figures in the formulæ for voltage gain given earlier, the gain $= \dfrac{9\cdot35 \times 2500}{1870 \times 2500\ (10\cdot35)} = 0\cdot845$.

Similarly the effective output impedance $= \dfrac{1870}{10\cdot35} = 181$ ohms and if the valve has a c_{ga} of $0\cdot7\ pF$ and a c_{in} of $11\cdot5\ pF$ then the effective input capacity $= 0\cdot7 + \dfrac{11\cdot5}{9\cdot35} = 1\cdot93\ pF$.

The distortion can be calculated by measurement of the lengths of EF and AD in Fig. 43, F being the centre of the length AD. The percentage of 2nd harmonic distortion is $\dfrac{EF}{AD} \times 100$ which equals $1\cdot93$ per cent. in this example. At point E the anode current is $56\cdot5$ mA so the anode-cathode voltage $= 500 - \dfrac{56\cdot5 \times 2500}{1000} = 359$ volts and the anode dissipation $= \dfrac{359 \times 56\cdot5}{1000} = 20\cdot25$ watts. The anode to cathode voltage at point A is obviously 500 volts and at point D $= 500 - \dfrac{122 \times 2500}{1000} = 195$ volts.

So that the power output $= \dfrac{(500 - 195) \times (0\cdot122 - 0)}{8} = 4\cdot65$ watts.

The above example provides all the necessary information to design a cathode follower output stage using a resistor as the output load. In the cases where the load is a tuned circuit or a loaded transformer the procedure is similar except that since the D.C. resistance of such a circuit is likely to be

very low, and unless the voltage drop in the resistance exceeds the normal grid bias, in which case a positive grid bias must be supplied, bias may be obtained by means of a resistor shunted by a large condenser in the cathode lead in the conventional fashion. The value of the autobias resistor as calculated is the value of the resistor plus any resistance present in the coil or transformer.

For a given valve it can be shown that the optimum load for minimum distortion is the same, whether the load is connected in the anode or the cathode, but in the latter case, *i.e.* the cathode follower, the distortion and output impedance are lower than in the former. It is also convenient at times to divide the load so that part of the load is in the anode and part in the cathode, this arrangement allows of a greater gain than as a true cathode follower, and less distortion and lower output impedance than when used as a conventional amplifier.

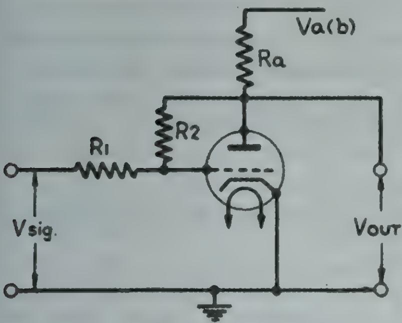


Fig. 44a.
Basic circuit of an anode follower.

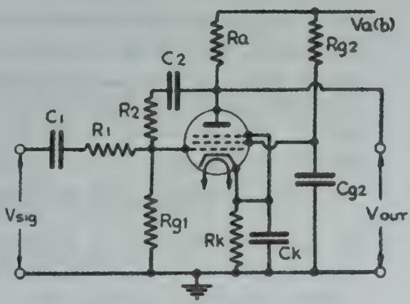


Fig. 44b.
Practical anode follower circuit.

Anode Followers

When it is desired to use a valve as a voltage amplifier with a wide frequency response together with a stable voltage gain so that it is independent of changes in supply voltages, valve replacements, etc., it is customary to apply a high degree of voltage feed back. When such feed-back is applied in a manner to be described the circuit is known as an anode follower.

The basic circuit is shown in Fig. 44a. Voltage feed-back is applied by injecting a known fraction of the output voltage V_{out} effectively in series with the input voltage, V_{sig}. The feed-back voltage V_f is obtained from V_{out} by means of the potential divider formed by R_1 and R_2 so that $V_f = \frac{R_1}{R_1 + R_2} \times V_{out}$. The actual input to the grid is $V_f + V_{sig}$, the feed-back being negative due to the phase reversal in the valve.

As the valve is to be used as a voltage amplifier it will be biased so that no grid current flows. V_{sig} will cause some current to flow through R_1, but since the grid takes no current it must also flow through R_2, and as the current is equal in each, if they are of the same value, the amplitude of the voltages will be equal. It is assumed that the gain of the valve is high, in other words the input to the grid is small. It follows, therefore, that the input

and output wave forms are equal in amplitude and the anode voltage follows the input signal. Two conditions are necessary that the above may be true. No current must flow from the junction of R_1 and R_2 either into the grid or through the grid leak, and the amplification of the valve must be high enough to ensure a grid amplitude small compared with V_{sig} or V_{out}. For this reason it is usual to employ a high slope tetrode or pentode. If R_1 and R_2 are not of equal value, but still have equal currents the amplitude of the voltage developed across them will be proportional to their resistance values. Thus if R_2 is greater than R_1 the output will be greater than the input; hence the amplification of the stage will depend on the ratio of the resistances and to a very large extent it will be independent of everything else. This amplification with feed-back is equal to R_2/R_1.

If the stage gain is n, then $R_2 = n \times R_1$. The values of R_1 and R_2 should be reasonably high compared with R_a. For example, if R_a is 500,000 ohms and a gain of 10 is required, R_2 should be 500,000 ohms, and R_1 50,000 ohms and this will load the preceding stage with 50,000 ohms. If a higher input impedance is essential R_2 should be increased to 2 megohms and R_1 proportionately to 200,000 ohms, giving an input impedance of 200,000 ohms.

The output impedance of the stage $= \dfrac{n + 1}{g_m}$ ohms where n is, as before, the gain with feedback, or R_2/R_1, and g_m is in amps per volt. It is evident that if n has a low value and the valve slope is high the output impedance is quite low. When R_1 equals R_2 the impedance is $2/g_m$ or twice that of a cathode follower.

The basic circuit in Fig. 44a cannot be applied as it is shown, because there is no provision for preventing the D.C. from the anode reaching the grid and an input with no D.C. path would leave the grid floating. A practical circuit for a pentode is shown in Fig. 44b, here it will be seen that the blocking condensers C_1 and C_2 and the grid leak R_{g1} have been introduced as well as the normal decoupled autobias resistor R_k and the screen decoupling components R_{g2}, C_{g2}.

The values of the condensers C_1 and C_2 would at first sight appear to need to be very large, but in fact this is not so and the frequency response will be flat providing $R_1\ C_1 = R_2\ C_2$. In most cases suitable values are from 0·01 to 0·1 μF.

The value of R_{g1} would also appear to require to be very high, but in fact may be made equal to R_2 because if $R_{g1} = R_2$ and the gain without feed-back is 100 then only 1/100th of the current in R_2 will flow in R_{g1}.

At low frequencies R_{g1} should not be less than the impedance of the feed-back arm $R_2\ C_2$, which increases with reduction in frequency.

As an example of the method of determining the operating conditions for such a circuit, although a high slope pentode would generally be used in practice, it will be convenient to use as an example the low slope pentode,

whose characteristics are shown in Figs. 3 and 4 of Chapter 2. The method described in that chapter can be applied to determine the values and gain without the feed-back. These values when determined can be applied to Fig. 44b as follows :—

With R_a 500,000 ohms voltage gain without feedback, A = 230, neglecting the succeeding valve grid leak, which is omitted in Fig. 44b.

Cathode resistor R_k = 3,800 ohms, screen resistor R_{g2} = 2·6 megohms, output voltage = 62 volts R.M.S. and slope at the operating point 0·46 mA/V. If it is assumed that a gain of 10 is required from the anode follower then the ratio of the gain without feed-back to gain with feed-back is 230/10 = 23 and R_2/R_1 = 10, so that if R_2 = 500,000 ohms, then R_1 = 50,000 ohms. The input impedance will be 50,000 ohms and the output impedance = $\frac{(10 + 1) \times 1000}{0 \cdot 46}$ = 24,000 ohms. The value of R_{g1} can conveniently be given a value of 500,000 ohms, so as not to be low compared with R_2. If C_1 is made 0·1 μF then C_2 should be 0·01 μF.

The voltage gain of an anode follower $= \dfrac{R_2}{R_1 + \dfrac{(R_1 + R_2)}{A}}$ where A is voltage gain without the feed-back circuit. In the example A is 230, so the gain is 9·7 which is approximately the figure of 10 mentioned earlier. If at either extremes of the frequency response the gain without feedback had fallen ten times, then A would be 23 and the gain $= \dfrac{500{,}000}{50{,}000 + \dfrac{550{,}000}{23}} = 6 \cdot 7$.

So that a fall in gain of ten times or 20 *db* without feed-back becomes approximately 1·5 times or 3·5 *db* with feed-back.

The circuit described is quite applicable to the use of a transformer in the output instead of a resistor R_a, the value of the load being the impedance of the transformer and the calculations of gain without feed-back are then performed as described in Chapter 2.

The resistances R_1, R_2 and condensers C_1 and C_2 may be very conveniently switched in such a way that their ratios bear a constant relationship. This arrangement is very suitable for an amplifier which requires several fixed positions of gain, *e.g.* for use with an oscilloscope. If, due to wiring stray capacities, there is some capacity across R_2 this may be balanced by a small condenser across R_1, or vice-versa, so that the time constant of each is the same. The frequency response will then be unaffected.

Phase Splitters

There are many circuits not employing a transformer that may be used for the purpose of obtaining a push-pull output from a single ended input. They all basically comprise either a single or double valve arrangement using resistance capacity coupling. It is proposed to describe only a few of the methods.

Paraphase Amplifier

In this arrangement two similar triodes are used, the grid of the second valve being driven from the first valve to reverse the phase of the output. The arrangement is shown in Fig. 45a. V_1 is a normal R.C. coupled voltage

amplifier, its output appearing across R_1, R_2. The part of this output which appears across R_2 feeds V_2 which acts also as an R.C. coupled voltage amplifier providing the opposite phase. The ratio of R_1 to R_2 is adjusted so that $\frac{R_1 + R_2}{R_2}$ is equal to the voltage gain of V_2. Then the output voltages of both valves will be equal and opposite if $R_1 + R_2 = R_3$. The operating conditions for each valve are calculated by the method described for voltage amplifiers, but the values of R_1, R_2 are critical and are chosen to suit the voltage gain. The common cathode resistor has a value of half the value for each valve alone. The by-pass condenser may be omitted with some loss of balance at the higher frequencies.

Split-load Phase Invertor

This type uses a single valve, having half the load in the anode and half in the cathode as shown in Fig. 45b. With respect to the load in the anode the valve acts as a normal voltage amplifier but with the load in the cathode it forms a type of cathode follower. Since the current in both loads is equal, the voltages are equal if $R_1 = R_2$ and $R_3 = R_4$, but because of the negative feed-back the gain is reduced to a value of approximately unity. Since the ratio of the feed-back voltage to the output voltage $= \frac{R_1}{R_1 + R_2}$ and since $R_1 \approx R_2$, the feed-back ratio is 0·5. In other words a V_{sig} of 1 volt will produce a 1 volt across both anode and cathode loads. Since there is current feed-back as far as R_1 is concerned the output impedance will be increased, but as far as R_2 is concerned the impedance will be decreased, so that the source impedance from which the push-pull valves will be working will be unbalanced. Hence this type of phase splitter should not be used if the push-pull stage runs into grid current at any time or is operated other than strictly Class A1. The operating conditions should be determined as if the valve was working as an R.C. coupled amplifier with R_2 omitted and R_1 twice the value to be used. The effective output voltage will be half the value so calculated.

Cathode Coupled Phase Splitter

Two similar triodes are used with the input applied to the grid of one triode only, the other grid being undriven. The output is taken from the anodes, and the coupling between the valves is by a common resistor R_b in the cathode lead. The circuit is shown in Fig. 45c. With this arrangement the output voltages are balanced in amplitude and impedance and the voltage gain is about a quarter of that obtained for each triode used in a normal R.C. coupled amplifier. The value of R_b should be approximately $\frac{R_a}{2}$. The operating conditions are determined for each triode by assuming an anode load resistance of $R_a + (2 \times R_b)$ and a cathode bias resistance of $2 \times R_k$.

Anode Follower Paraphase Amplifier

If an anode follower is arranged so that it has a high degree of feed-back,

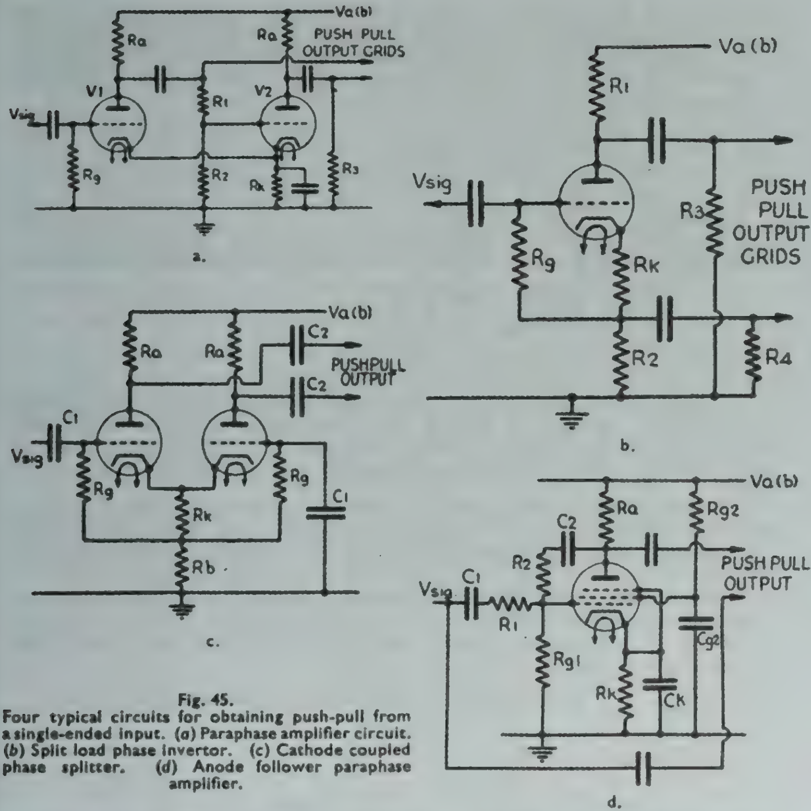


Fig. 45.
Four typical circuits for obtaining push-pull from a single-ended input. (*a*) Paraphase amplifier circuit. (*b*) Split load phase invertor. (*c*) Cathode coupled phase splitter. (*d*) Anode follower paraphase amplifier.

then the gain will be unity and the output will be equal in voltage but opposite in phase to the input. A circuit is shown in Fig. 45d. Referring back to the basic anode follower circuit, if $R_1 = R_2$ and $C_1 = C_2$ and if the voltage gain without feed-back is very high, then the gain $= 1$. The outputs have an impedance differing to some extent because one output has a value equal to the source impedance of V_{sig}, and the other is equal to $2/g_m$. The circuit constants, etc., are derived as for an anode follower.

Grounded Grid Amplifiers

Increasing use of higher frequencies has led to the development of a circuit in which the grid of a triode instead of the cathode is grounded. The basic arrangement is shown in Fig. 46 and it is to be noted that the grid is common to both input and output circuits. This has advantages over a conventional grounded cathode because the inter-electrode capacity common to the input and output circuits and which must be neutralised for stability is now the capacity between anode and cathode or filament, and with valves of modern

construction the value of this capacity is sufficiently low to permit a smaller neutralising condenser or frequently none at all, even at very high frequencies. The absence of such condensers reduces the limitation which this capacity imposes on the peak power output and useful anode voltage swing for a given bandwidth.

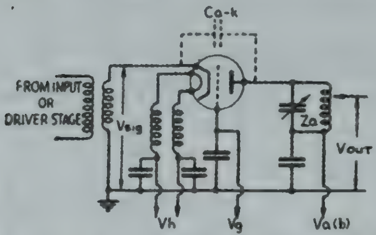


Fig. 46.
Basic circuit of a grounded-grid amplifier.

Since the driver stage feeding the cathode is effectively in series with the output load, some of the driving power appears in the output as useful power, but for this reason it is not usually possible to modulate a grounded grid amplifier alone, and some modulation must be applied to the driver stage.

When a grounded grid amplifier is used in a receiver as a voltage amplifier, the chief advantage is the higher gain achieved up to about 2,000 Mc/s. with a much improved signal to noise ratio, compared with conventional circuits.

As the input is applied to the cathode, degeneration is present and the input impedance is therefore low.

If Z_a is the output load impedance, then the input impedance

$Z_{in} = \dfrac{Z_a + r_a}{1 + \mu}$ the values of r_a and μ being at the operating point.

$$\text{The input power} = \frac{(V_{sig})^2 (1 + \mu)}{Z_a + r_a}$$

$$\text{The output power} = \frac{(1+\mu)^2 (V_{sig})^2 \times Z_a}{(r_a + Z_a)^2}$$

The power gain is thus proportional to $\dfrac{Z_a}{Z_{in}}$ *i.e.* the ratio of the load impedance to the input impedance, so that a grounded grid amplifier can be regarded as an impedance transformer with a step-up ratio which is the opposite case to that of a cathode follower.

The cathode being at R.F. potential implies that the heater or filament voltage should be supplied via R.F. chokes or by an inter winding on the input coil. It is generally undesirable in using an indirectly heated cathode to dispense with these R.F. chokes, as this would place the heater-cathode insulation across the input voltage.

The optimum performance is to be obtained with valves especially designed for use as grounded grid amplifiers but quite useful performance may be obtained with valves designed for normal purposes, providing they have

low capacities, such as are afforded by use of an all glass base (button B7G), and have a reasonably high amplification factor μ. Pentodes may be connected as triodes for this purpose but in this case it is the μ as a triode which is relevant.

The gain and bandwidth obtained with grounded grid amplifiers is frequently disappointing. This is because due regard has not been paid to matching of the impedance both input and output. Particularly when used as a voltage amplifier, simple tuned circuits between stages generally result in a poor gain for a given bandwidth and a properly designed filter is preferable.

Fig. 47 shows a typical circuit of a two-stage grounded grid pre-amplifier. The input is applied to the cathode of V_1 via an input coil L_1 which has the heater supply interwound with the main winding. The output of V_1 is fed to the filter circuit comprising L_2 and L_3, L_3 also having an interwound heater supply. The low impedance output across L_3 drives the cathode of V_2. The output of V_2 is fed to a filter circuit comprising L_4 and L_5, the output being coupled to L_5.

The bias resistor for the two valves is R_3 decoupled close to each coil by C_3 and C_7. The heaters are by-passed by C_1, C_2, C_6 and C_8. The anode voltage is decoupled by R_1, C_4 and R_2, C_{10}, and fed to the anodes via R.F. chokes. All the coils are tuned by adjustable dust cores or metal plungers, and have no tuning capacity across them other than the stray capacities.

The operating conditions, such as grid bias, etc., are derived from the characteristic curves in a similar manner to that used for the same valve employing a normal grounded cathode.

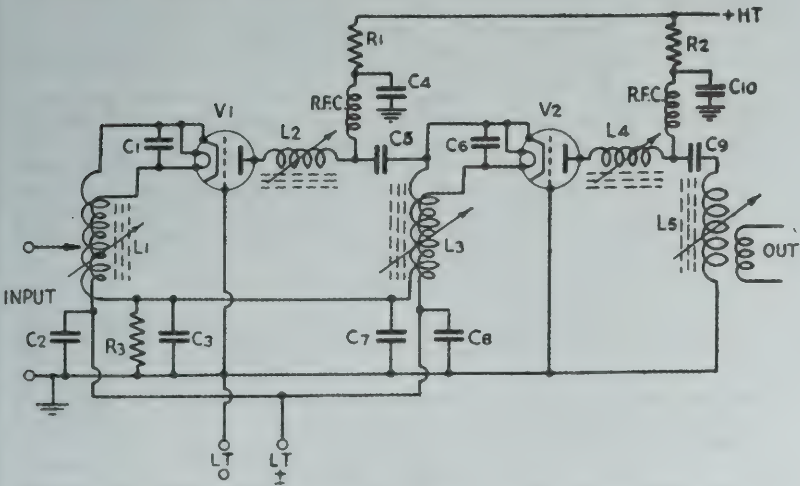


Fig. 47.
Typical circuit of a grounded-grid U.H.F. pre-amplifier. Gain approximately 20 db with 6 Mc/s. bandwidth at frequencies from 45-200 Mc/s.

CHAPTER 9 NOISE GENERATED IN VALVE AMPLIFIERS

MODERN designs of communication equipment frequently require the use of wider band widths or higher frequencies than were customary in the past, and in consequence the problems of both valve and circuit noise have become more prominent.

It is therefore considered that the inclusión in this book of the principles of noise calculations, even though they are primarily a circuit matter, is justified.

In any valve amplifier there are three primary sources of noise (*a*) Thermal agitation or "Johnson" noise in the circuits connected with the valve. (*b*) Valve or "Shot Noise." (*c*) Induced grid noise.

"Johnson" Noise

This is due to the fact that when current flows through any conductor, the electrons in the conductor do not move in a uniform steady manner, because the heat in the conductor causes agitation of the molecules of the conductor material giving rise to an unsteady movement. This unsteadiness causes a voltage drop across the conductor proportional to the current and to the resistance, and since the voltage is caused by random motion the voltage constitutes a random noise voltage and, therefore, occupies an infinite frequency band.

The value of the mean square noise voltage is given by:

$$4KTR\Delta f$$

where K = Boltzmans constant (Joules per degree Kelvin) = $1 \cdot 37 \times 10^{-23}$
T = Absolute temperature in degrees Kelvin;
Δf = the bandwidth in cycles per second;
R = value of the resistance in ohms or the dynamic resistance in the case of a tuned circuit.

The bandwidth Δf is the effective bandwidth or the area under the curve of the amplitude of the power versus the frequency, but for practical purposes little error results from taking Δf at the points on the curve where the power has fallen to one half or the voltage to $0 \cdot 707$ of its maximum value.

To take an example, consider a resistor of 100,000 ohms at normal room temperature connected to the input of an A.F. amplifier of having a bandwidth of 30 — 10,000 cycles. Δf is approximately 10,000, T is 293°K (273° + 20°), so that the noise voltage

$$= \sqrt{4 \times 1 \cdot 37 \times 293 \times 10000 \times 100000 \times 10^{-23}}$$
$$= 4 \text{ microvolts.}$$

Taking a further example of a television receiver designed for a video bandwidth of 3 Mc/s. and having the first tuned circuit loaded with 2,500 ohms, the dynamic resistance of the tuned circuit being assumed to be high compared with this figure. The total band width, assuming a double side band system, is 6 Mc/s.

The noise voltage from the input circuit is :

$$= \sqrt{4 \times 1 \cdot 37 \times 293 \times 6 \times 10^{6} \times 2500 \times 10^{-23}} = 15 \cdot 5 \text{ microvolts.}$$

Fig. 48 shows a chart calculated from the formula given above for a normal room temperature, from which the value of the noise voltage may be found without calculation.

Shot Noise

This form of noise is generated in all electronic valves by random fluctuations in the anode current, and is so called because when amplified it sounds as if shot were falling on a metal surface. Although the noise is generated in the anode circuit, it is more convenient to refer to it as the noise which would be generated by an equivalent noise voltage applied to the grid. This process is quite simple because $\frac{I_a}{g_m} = V_g$. As the "Johnson" noise is calculated from the resistance value, it is in fact more convenient still to quote the Shot Noise not as a grid voltage but as an equivalent resistance (R_{eq}) which connected in series with the grid would generate the same value of noise.

The formula is the same as for "Johnson" noise but R_{eq} is substituted for R. Knowing the values for R_{eq} it is possible to compare the relative merits of different valves regardless of the bandwidth, whereas if the noise voltage were quoted the bandwidth would necessarily have to be stated for each case.

The value of R_{eq} is concerned with the mutual conductance of the valve

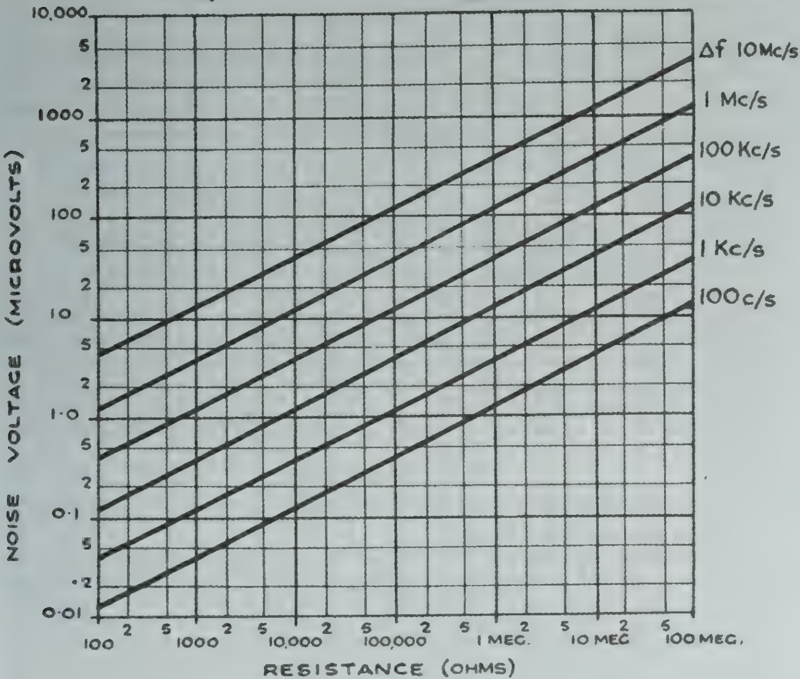


Fig. 48.
A chart for determining the noise voltage developed in different values of resistance for various bandwidths Δf at normal room temperature of 20°C.

and the total cathode current I_k. For pentodes the value is higher than triodes because the anode receives only a portion of the total cathode current and multi-grid valves such as frequency changers have a still higher value of R_{eq}, due to the poor ratio of anode current to total cathode current.

The values of R_{eq} in kilohms may be calculated as follows knowing the mutual conductance (g_m) or conversion conductance (g_c) in milliamps/volt at the working point.

For triodes used as amplifiers, $R_{eq} = \dfrac{2 \cdot 5}{g_m}$

For triodes used as frequency changers, $R_{eq} = \dfrac{2 \cdot 5}{g_c}$

For pentodes used as amplifiers, $R_{eq} = \dfrac{I_a}{I_a + I_{g2}} \left(\dfrac{2 \cdot 5}{g_m} + \dfrac{20\, I_{g2}}{{g_m}^2} \right)$

For pentodes used as frequency changers, $R_{eq} = \dfrac{I_a}{I_a + I_{g2}} \left(\dfrac{2 \cdot 5}{g_c} + \dfrac{20\, I_{g2}}{{g_c}^2} \right)$

For multi-grid frequency changers, $R_{eq} = 20 \dfrac{I_a (I_k - I_a)}{I_k \times {g_c}^2}$

For example the pentode valve used in Fig. 3 employed as an amplifier has an anode current of $2 \cdot 3$ mA, and screen current of $0 \cdot 7$ mA, and g_m of $1 \cdot 2$ mA/V at a grid voltage of –3, so the value of $R_{eq} = 9 \cdot 1 \times 1{,}000$ ohms, and the noise voltage for a 10 kc/s. bandwidth at room temperature 20°C. will be $1 \cdot 2$ microvolts.

Induced Grid Noise

There is a second form of noise present in a valve which is induced into the grid structure by the random fluctuations in anode current. The induced voltage on the grid is affected by the nature of the external impedance between grid and cathode. The input impedance of a valve comprises a resistive and reactive term. The resistive term is composed of the transit time effect and the inductance of the cathode lead. Both these are of negligible magnitude at low frequencies but at high frequencies become comparable with or lower in value than the external grid circuit impedance. The reactive term is that due to the capacitance and is not responsible for any noise. The induced grid noise for a valve having the control grid adjacent to the cathode is given by :

$$1 \cdot 4 \times 4 \times KTR_e \Delta f,$$

where R_e is the resistive component of the input impedance, and T is the Absolute cathode temperature.

The cathode temperature of most indirectly heated receiving valves can be taken as about five times the normal room temperature in degrees Absolute.

The input impedance of valves can be measured, and for valves intended for V.H.F. operation this figure is usually available from the manufacturers. If the cathode lead lengths are very short or two or more cathode leads are employed, it may be assumed that the figure quoted is the value R_e. This value varies quite considerably with frequency and may be neglected at frequencies below 10 Mc/s. Above this figure the value can be taken as

50,000 ohms at 20 Mc/s., 10,000 ohms at 50 Mc/s. and 2,000 ohms at 100 Mc/s. as a first approximation.

Calculation of Total Noise

Fig. 49 shows the equivalent circuit of a receiver input. The aerial will contribute some noise voltage depending upon its radiation resistance. If the coupling between the aerial and the grid involves a step up or down due allowance must be made for this by multiplying the aerial noise voltage by the turns ratio.

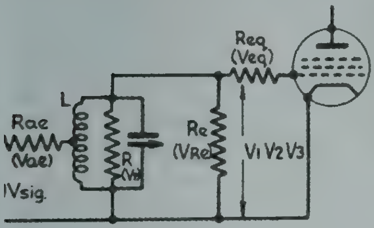


Fig. 49.
Equivalent circuit of noise resistances and noise voltages present in the input of a typical receiver.

R_{ae} = Aerial radiation resistance.
R = "Johnson" noise in the tuned circuit.
R_e = Induced grid noise.
R_{eq} = Shot noise.
V_{ae} = Aerial noise voltage.
V_r = Noise voltage due to the tuned circuit.
V_{re} = Noise voltage due to induced grid noise.
V_{eq} = Noise voltage due to shot noise.
N = Turns ratio of aerial coil L.

As the noise produced by each source is a random effect and is calculated on a power basis, the separate components cannot be added directly, but the total must be obtained by taking the square root of the sum of the squares, thus:

$$\text{Total noise} = \sqrt{V_1^{\,2} + V_2^{\,2} + V_3^{\,2} + V_4^{\,2} + \text{etc.}}$$

The voltages V_1, V_2, etc., are not, in fact, the calculated noise voltages, because if Fig. 49 is examined it will be clear that the noise voltage generated by the aerial, after being stepped up by the ratio of the aerial coil, is in parallel with R and R_e and these resistances absorb some of the noise and it does not, therefore, all reach the grid of the valve. This applies also to the other noise voltages.

Referring again to Fig. 49, in order to find the noise voltage of the aerial as applied to the grid, if V_{ae} is the noise voltage generated by the aerial, calculated by using the value R_{ae} in the "Johnson" noise formula, and N is the step up turns ratio of the coil, then the value of V_1, the effective value applied to the grid of V_{ae} is calculated as follows:

$$V_1 = \frac{V_{ae} \times N}{(R_{ae} \times N^2 + \dfrac{R \times R_e}{R + R_e}} \times \frac{R \times R_e}{R + R_e}$$

Similarly V_2 is the effective value of V_r.

$$V_2 = \frac{V_r}{R + \dfrac{R_{ae} \times N^2 \times R_e}{(R_{ae} \times N^2) + R_e}} \times \frac{R_{ae} \times N^2 \times R_e}{(R_{ae} \times N^2) + R_e}$$

and V_3 is the effective value of V_{re}

$$V_3 = \frac{V_{re}}{R_e + \dfrac{R_{ae} \times N^2 \times R}{(R_{ae} \times N^2) + R}} \times \frac{R_{ae} \times N^2 \times R}{(R_{ae} \times N^2) + R}$$

To take a practical example, consider Fig. 49 to be a stage of the television receiver referred to earlier. If the aerial is a dipole of 70 ohms (R_{ae}), and the receiver band width is 6 Mc/s., V_{ae} at room temperature calculated by the "Johnson" noise formula is 2·6 microvolts. The noise due to the tuned circuit, if this is loaded with 2,500 ohms (R) as before, was earlier calculated to be 15·5 microvolts so that this is the value of V_r. If the first valve is an R.F. pentode having a mutual conductance of 7·5 mA/V, an anode current of 10mA and a screen current of 2·5 mA, then the "Shot" noise equivalent resistance R_{eq} is 970 ohms, and for the same bandwidth using the same formula the value of V_{eq} will be 9·7 microvolts. If the valve has an input impedance (R_e) of 6,000 ohms at 45 Mc/s. then the value of induced grid noise voltage V_{re}

$$= \sqrt{1{\cdot}4 \times 4 \times 1{\cdot}37 \times (5 \times 293) \times 6 \times 10^6 \times 6000 \times 10^{-23}}$$
$$= 63{\cdot}5 \text{ microvolts.}$$

The effective value of the noise applied to the grid due to the aerial if the turns ratio N of the aerial coil is 5 : 1

$$= V_1 = \frac{2{\cdot}6 \times 5}{(70 \times 5^2) + \dfrac{2500 \times 6000}{2500 + 6000}} \times \frac{2500 \times 6000}{2500 + 6000} = 6.5 \text{ microvolts.}$$

Similarly

$$V_2 = 5{\cdot}5 \text{ microvolts}$$

and $V_3 = 9{\cdot}3$ microvolts.

The Shot Noise voltage must then be added to the above voltages. This voltage is not affected by the resistance in parallel because it is considered as a noise voltage in series with the grid.

Thus the total noise voltage

$$= \sqrt{9{\cdot}7^2 + 6{\cdot}5^2 + 5{\cdot}5^2 + 9{\cdot}3^2} = 15{\cdot}9 \text{ microvolts.}$$

In a receiver it may happen that there is considerable noise produced by the second stage if the first happens to be an R.F. stage and the second a frequency changer. In this case the noise generated by the second stage comprising R, R_e and R_{eq} is totalled, and referred back to the grid of the first stage by dividing the value obtained by the gain of the first stage, when it is then added to the four sources of noise already present, making a fifth term in the total noise formula.

The example worked out indicates that the value of total noise is the result of various factors and reducing one alone would not reduce the noise very much. Although the induced grid noise is quite a large figure its effective value is so reduced by the other parallel resistances that it does not contribute nearly as much to the total noise as might be expected.

CHAPTER 10 VALVES FOR V.H.F.

NORMAL TYPES OF VALVES FOR OPERATION AT V.H.F.

Frequency Limitation

THE main limitations to the operation of valves at very high frequencies are due to lead inductance, inter-electrode capacities and transit time. These limitations apply whether the valve is used as an amplifier or as an oscillator. Oscillation will be possible at a slightly higher frequency than the limiting frequency for useful amplification, but the difference is small.

The effects of lead inductance and internal capacities are mainly troublesome because they occur together; if either factor could exist alone it would not be so difficult to manage. The capacities, if they existed alone, would simply form part of the tuning circuits. At high frequencies the reactance of these capacities falls, and the capacitive current increases and may assume a considerable magnitude. This current has to flow through the lead inductance, the reactance of which will be increasing as the frequency is raised, and the H.F. ohmic resistance of the lead which will also increase with frequency. The result is that the loss in the ohmic resistance is increased and the internal grid voltage is not in phase with that supplied by the external circuit. An apparent phase displacement also arises from the finite time of electron flight, with the result that any change in the density of the electron stream will not be felt at the anode until a moment later. If this delay becomes an appreciable part of a cycle, as it may at V.H.F., it is tantamount to a phase shift. These phase shifts and losses additively introduce so much circuit damping that amplification and oscillation become impossible.

It follows, therefore, that in order to achieve satisfactory operation at the higher frequencies these effects must be reduced as far as possible. Lead inductance and resistance can be reduced by use of short thick electrode leads, the highest frequencies being reached by making the electrode and its lead into one component, which becomes a part of a resonator system, such as long line or cavity. Alternatively two or more parallel connections may be made to the same electrode either to separate terminals or to one terminal. Cathode lead inductance is particularly troublesome because it is common to both grid and anode circuits. Where twin cathode leads are brought out to separate terminals they are used to isolate these paths as shown in Fig. 50.

To reduce transit time it is necessary, for a given anode potential, to reduce the anode to cathode clearance, which in turn will raise the capacities unless the areas of the electrodes can be proportionately reduced. The smaller electrode areas are desirable even if no reduction of clearance is considered, but the reduction in area, particularly of the anode, may impose a severe dissipation limitation. A reduction in transit time is also possible by raising the anode potential, but as the time of flight is proportional to the inverse of the square root of the voltage the latter becomes prohibitively high before any major decrease has been achieved.

"Television" Type Pentodes

In order to achieve the wide bandwidths needed for television I.F. amplifiers it is necessary to introduce resistance damping in the tuned circuits; it is obviously not worth while going to a lot of trouble to reduce valve damping and the first attempts to raise the frequency of operation were made simply by making high slope versions of normal valves. The higher slope

demands larger cathodes and grids with reduced grid to cathode clearance so that the input impedance is very low. Unless associated with some form of construction which enables a marked reduction in lead loss to be achieved, these valves are not very useful, except in wide band amplifiers. The frequency limit is about 50 Mc/s.

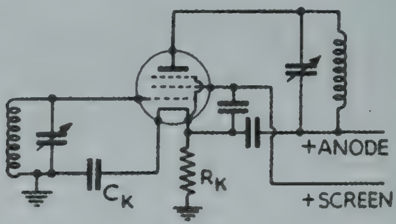


Fig. 50.
Arrangement of a valve, with two cathode connections, to ensure isolation of anode and grid circuits. R_K is the normal bias resistor and C_K is the bias by-pass condenser.

" Acorn " Valves

These are miniature editions of normal electrode structures mounted directly on thick electrode leads which come straight through the glass wall to connect to the circuit. The miniature construction results in a considerable reduction of capacities and transit time and the lead inductance and resistance is also greatly reduced. The valves will operate most successfully up to about 500 Mc/s., the only troubles being the very limited dissipation and the high manufacturing costs.

" All Glass " or " Ring Seal " Valves

A very great improvement in performance of valves of more or less normal construction has been achieved by abandoning the pinch type of seal in favour of the ring seal method.

The electrodes are mounted directly on short pins passing through a glass disc, the pins themselves forming the contacts for insertion into the valve holder. The bulb is afterwards placed over the mounted electrodes and united with the edge of the glass disc. No great reduction of capacities within the electrodes themselves is achieved but the additional capacity existing between the leads of a pinch type valve has been almost eliminated and the lead inductance and resistance has been considerably reduced. The majority of the American " all metal " valves fall into this class as of later years they have used the glass disc construction, the edge of the disc being sealed to the metal shell. The leads are not stiff enough to act as contacts, so they are connected to pins inserted into the base wafer. The true " all glass " valves are the American Loctal (B8G) and the British B9G ranges. The upper frequency limit of these valves is about 150 Mc/s.

A recent development has been the use of miniature valves of all glass construction on a seven pin base (B7G). The American "9000" series are direct equivalents of the " acorns " with this form of construction. Their frequency limit is not quite so high as the true acorns, but they are easier to manufacture and are, therefore, cheaper. A number of British and American valves have been introduced using this base and both battery and mains versions are available. They are effective up to 300 Mc/s., but there is a large variety of types and this figure is only a very rough guide.

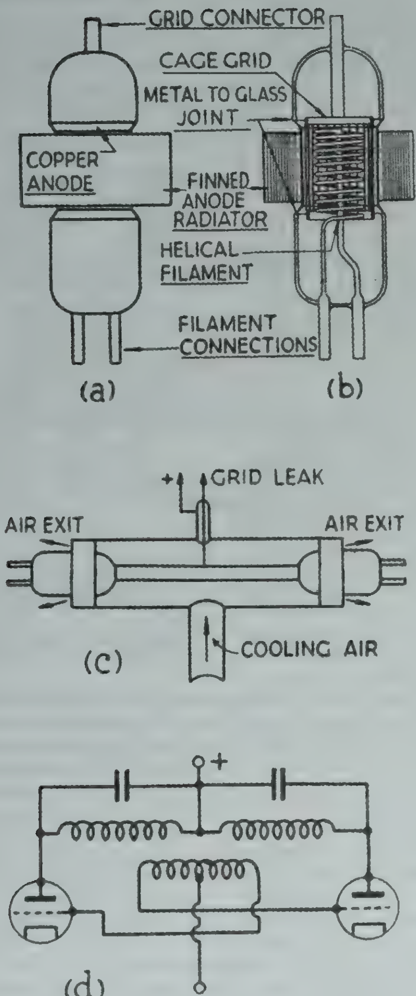


Fig. 51.

Concentric construction of a transmitting valve. (*a*) External appearance. (*b*) Cross-section. (*c*) Concentric line oscillator. (*d*) Equivalent circuit of Fig. 51c.

" Concentric " Valves

Transmitting valves capable of up to 200 watts dissipation, with forced air cooling, have been developed to operate up to approximately 250 Mc/s. They are intended to work with tuned circuits consisting of concentric line tuning units and they are built to form part of these lines. A sketch of a valve of this type is shown in Figs. 51a and 51b, while Fig. 51c shows one form of the concentric line tuning system.

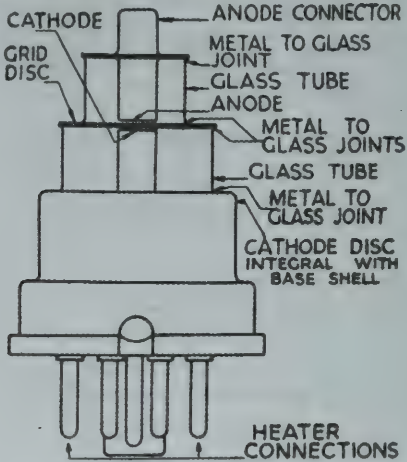


Fig. 52.
Planar Triode valve or " Lighthouse " tube.

Disc or Planar Valves

This form of construction, although very costly, has permitted the effective range of the ordinary negative grid valve to be extended well into the micro-range region. The amateur 2,300 Mc/s. band is just about within their reach. Fig. 52 shows the method of construction. Their appearance has given rise to the popular name of " lighthouse " tubes. It will be noted that the effective cathode and anode areas are very small, the former demanding a very high emission in terms of milliamperes per square centimetre, and the latter restricting the dissipation considerably. The valves are intended to be used with cavity resonators as tuning elements, and to achieve the highest possible frequencies the valves themselves must become a part of the resonator system. Fig. 53 shows a grounded grid version of the tuned plate-tuned grid oscillator built with cavity resonators.

It is interesting to note that the shielding action of the grounded grid so isolates the anode and cathode circuits that, for oscillation, external feed-back has to be provided by means of inter-connected loops, the equivalent diagram looking like the familiar link coupling.

SPECIAL VALVES FOR MICROWAVES

Cavity Magnetrons

The old type of magnetron oscillator is a diode with a cylindrical and concentric cathode and anode, the anode being split into two or more segments,

placed in a magnetic field parallel to the cathode. Tuned circuits are connected across the segments of the anode, or between the anode and cathode.

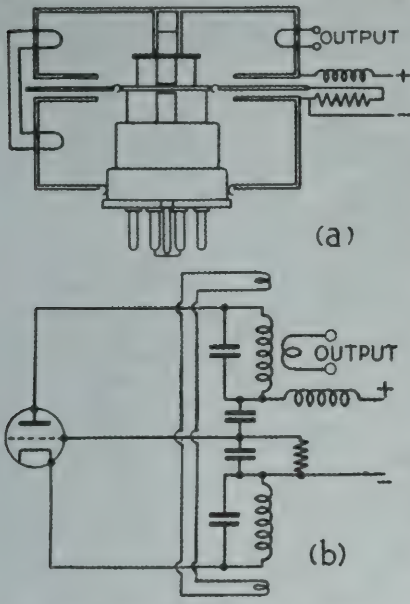


Fig. 53.
Planar Triode in resonant cavity system.
(*a*) Physical arrangement. (*b*) Equivalent circuit.

A recent modification of the magnetron utilises tuned cavities in place of the anode segments, so that the tuned circuit is built into the valve itself. The anode assembly is made from a solid copper block or built up from punched laminations, the internal construction taking the form of a series of holes arranged round the internal periphery of the block, each hole being coupled to the cathode region in the centre by means of narrow slots (Fig. 54). The number of holes increases with the frequency of operation (being usually 8 for a 10 cm. valve and 14 for a 3 cm. one), the R.F. output being taken from one of them *via* a coupling loop. The magnetic field is almost invariably provided by a permanent magnet placed externally to the anode block so that the field is at right angles to the ring of cavities, *i.e.* parallel to the cathode. In some designs the magnet is an integral part of the valve structure ("package" magnetrons) and cannot be separated.

The mode of operation of the valve is similar to that of the older type, which is already well covered in the literature, but a very brief description is as follows. Due to the axial magnetic field the path of the electrons leaving the cathode is curved, and for a certain value of field the electrons fall short of the anode and return to the cathode, travelling in circular orbits. The time taken for an electron to leave the cathode, travel round and return, determines the frequency of oscillation, the cavities being tuned to this frequency. Energy associated with the moving electrons will be given up to the space round the cathode whence it will be transferred to the tuned cavities and picked up by the coupling loop.

The valve may operate in different modes, the change from one to another being characterised by a sudden change in frequency. By "strapping" alternate cavities, *i.e.* connecting equipotential points by a copper link so that alternate cavities oscillate in phase, much more stable operation is achieved and the output is increased.

By means of this type of construction the cavity magnetron may have an efficiency as high as 65–70 per cent. and under C.W. conditions up to 100W output may be obtained at 1 cm. wavelength. A magnetron is not conveniently modulated in the ordinary way, although modulation be effected on the anode, and the valve is usually operated under pulse conditions. The H.T. voltage is applied in the form of pulses of the order of a few microseconds

in duration and of high energy content so that the valve is working only for the very short time of application of the pulse. Enormous peak powers have been obtained by pulse modulation, several megawatts being common in radar applications. It is the pulsed cavity magnetron which made possible centimetric radar and the valve is pre-eminent as a high power generator of microwave frequencies.

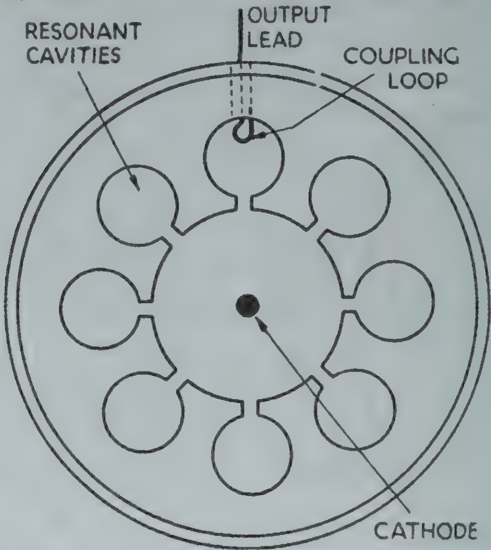


Fig. 54.
Anode block of a cavity magnetron

Velocity Modulation

By passing an electron stream through special electrode structures (Fig. 55), to which is applied a R.F. field, the stream may be velocity modulated, as parts of it will be speeded up and parts slowed down, due to the action of the field. Energy in the R.F. field is thus transferred to the electrons. By means of suitable devices this velocity modulation may be changed to amplitude modulation and the beam energy transferred to an output circuit.

Three different conversion methods are possible, (1) by deflecting the beam, (2) by applying a retarding field, and (3) by allowing the beam to drift in a field-free space. A magnetic field at right angles to the beam will effect deflection conversion since the slower moving electrons will be deflected through a larger angle than the faster moving ones and the main beam will split up into two smaller charge-density modulated beams. An electric deflecting field may also be employed. Retarding field conversion is done by allowing the electron stream to approach a negative potential electrode. Only the faster moving electrons will have sufficient kinetic energy to continue and be collected at the electrode, the slower ones being retarded and ultimately returned along their own path. This reflected stream is charge-density modulated and will supply R.F. energy to an electrode system placed in its path.

If a velocity modulated beam is allowed to pass through a field-free space (known as a " drift space "), the faster moving electrons will begin to overtake the slower ones, so that at some point along the beam alternate regions of high and low electron density will exist, *i.e.* the beam has become bunched and is now charge-density modulated at this point. The length of the drift space to produce bunching will depend on the magnitude of the radio frequency voltage existing across the electrode structure which is used to produce the velocity modulation. Only a very small voltage is necessary to produce deep amplitude modulation, provided that the drift space is long enough, since the electrons will in time sort themselves out into bunches even if there is only slight velocity modulation when the beam leaves the modulating electrodes. However, very long drift spaces are not a practical proposition since space charges may be set up in them which have a deleterious effect on the bunching action.

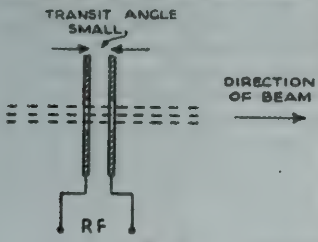


Fig. 55.
Velocity modulating grid.

It has been shown that at some point in the drift space the beam has become charge-density modulated, but the electrons in the bunches have different velocities and if the beam were allowed to continue to drift in field-free space they would un-sort themselves after a certain time and again bunch themselves. Thus in a long velocity modulated beam passing through field-free space there will be several equally spaced regions of charge-density modulation along the beam, the maxima being spaced by a transit angle of 2π radians. To abstract the energy from the beam it is only necessary to arrange for the output circuit to be placed at a point of maximum charge-density modulation.

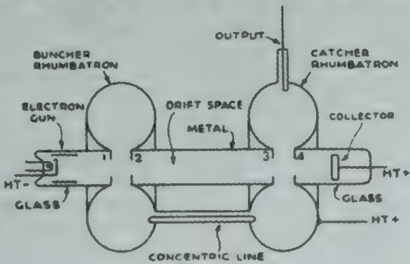


Fig. 56.
Klystron oscillator.

Double Resonator Klystrons

The klystron valve is a particular case of velocity modulation conversion by means of a drift space and is one of the more successful types. Fig. 56 shows such a valve. It contains an electron gun similar to that in a cathode ray tube, from which a narrow beam of electrons is projected along the axis of the tube and focused so that it will pass through the small apertures 1, 2, 3 and 4 across which the oscillating circuits are connected. The resonant circuits are of the cavity type and consist of hollow toroidal chambers, known as " rhumbatrons." The oscillating currents in this type of circuit flow on the inner surface of the chamber and there will be a high radio-frequency voltage across the neck 1, 2.

Suppose that the valve is oscillating, and ignore for the moment the length of concentric line connecting the two rhumbatrons. The electron beam from the cathode, accelerated by D.C. potential of perhaps several thousand volts applied to the body of the tube, will pass through the aperture 1, 2, across which there is a R.F. potential. This will cause velocity modulation of the

beam in the manner described. The velocity modulated beam then enters the field-free space between the two rhumbatrons, the drift space, where the faster electrons will begin to overtake the slower ones and the beam will become bunched. It should be noted that bunching in a drift tube really amounts to electron focusing in space and time. The distance between grid 2 and grid 3 is so arranged that the beam is bunched and de-bunched several times prior to reaching this grid. If we now assume that when the bunches of electrons arrive at the second rhumbatron (the catcher) the electric field across the neck 3, 4, is adverse, the electrons will be decelerated in their passage and their lost energy is imparted to the field, and hence to the catcher rhumbatron. The electrons are retarded almost to zero velocity in their passage through the gap 3, 4, and are removed by a collector electrode which may be at the same potential as the body of the tube.

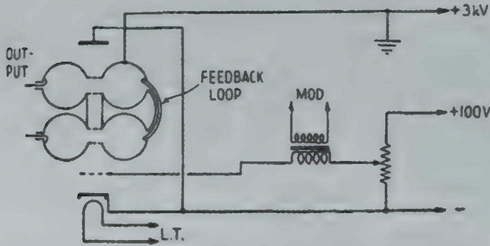


Fig. 57.
Power connections of a klystron.

If oscillation is to be maintained the beam must be bunched when it enters the catcher rhumbatron and the phase of the R.F. voltage across the neck of the catcher must be such as to retard the bunches of electrons entering the aperture. This latter is easily arranged by feeding energy from the catcher to the buncher rhumbatron by means of a short length of concentric line, the length of which is varied so as to establish the correct phase relationship between the voltages across the two rhumbatron necks. Radio frequency energy is taken from the catcher rhumbatron by means of a short length of concentric line connected to a coupling loop which is inserted in the rhumbatron, and which couples to the magnetic field round the axis.

Modulation is usually effected by applying speech voltages to one of the focusing grids and thus varying the intensity of the electron beam passing through the necks of the rhumbatrons, although other methods are possible and in certain cases give superior results. Fig. 57 shows a typical circuit for the operation of a klystron valve; it will be noted that as is common in cathode ray tube work, the positive side of the H.T. supply is earthed. This is because the external metalwork of the valve is connected to the rhumbatrons and would otherwise be at a high potential. By suitable connections the klystron may be used as an amplifier, mixer or detector as well as an oscillator and thus have advantages over some other types of microwave valves. When the klystron is used as an amplifier the length of concentric line connecting the rhumbatrons is removed and the signal to be amplified is fed into the buncher rhumbatron. In the case of valves intended only for oscillator work, a permanent feedback line between buncher and catcher can be used, and such a line is frequently built-in the valve so that although it has two resonators there is only one external connection, that for the output. Klystrons are not now often used for transmitting work, except sometimes as frequency multipliers, and certainly in all high power work have been

superseded by the cavity magnetron. They retain, however, a certain usefulness in receiver and instrument work and in low power applications.

Reflector Klystrons

This type of oscillator is rather similar in principle to the double resonator klystron but employs only one resonant circuit or rhumbatron and is intended mainly for use in microwave receivers where only a small amount of power is required. Fig. 58 shows the arrangement of a typical valve. The electron beam is focused so as to pass through the aperture in the rhumbatron and is thus velocity modulated on leaving. It then approaches a reflector electrode which is at a potential near that of the cathode, and may be either slightly positive or negative to the cathode. If it is at the cathode potential or slightly more positive, the velocity modulated beam, after leaving the modulating grid, enters a negative potential gradient and the part of the stream which possesses average, or more than average, kinetic energy will reach the electrode and give rise to reflector current. Part of the stream with kinetic energy below average will be retarded and turned back towards the cathode. This part of the stream is intermittent and is charge-density modulated.

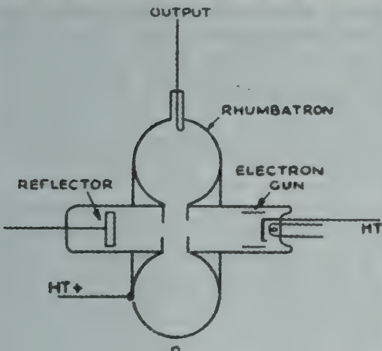


Fig. 58.
Arrangement of a typical reflector klystron.

Thus for oscillation to be maintained it is only necessary to arrange that the returning bunches of electrons are retarded during their second passage through the rhumbatron. If the reflector is operated with a potential negative to the cathode, no reflector current will flow and the entire beam will be reflected. In this case the returning beam is velocity modulated and becomes bunched when it enters the rhumbatron for the second time.

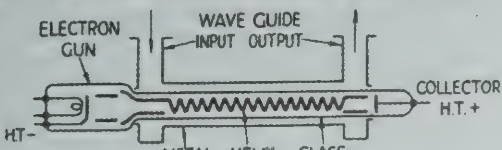


Fig. 59.
Travelling wave valve.

The reflector klystron is a very convenient valve and although it is obvious that it can be used only as an oscillator, it possesses considerable superiority over the double-resonator type for low power work. The reflector distance may be much shorter than the corresponding drift space in the double resonator valve due to the double transit of the electron beam and to the low velocity of the beam over part of the travel. There is only one resonator to tune compared with the two of a double resonator valve, which must be kept simultaneously and accurately in tune if any efficiency is to be realised. In other words, the buncher and catcher are always tuned to the same frequency as they are the same resonator. This advantage becomes peculiarly

apparent when the trouble of initially setting up a double valve has been realised. Reflector valves have attained a wide popularity as local oscillators in microwave receivers, due to this ease of tuning and their smallness and cheapness of construction.

Travelling Wave Valves

The basic principle of the magnetron oscillator is the interaction between a travelling electric field (rotating around the anode block) with an electron stream travelling at about the same order of velocity and this principle has been applied to a particular type of microwave amplifying valve. In the klystron the signal is used in the form of a stationary electric field applied to a rhumbatron, but the travelling wave valve, as implied in the name, uses the signal in the form of a travelling electric field which exchanges energy with the electron stream of the valve. The electric field travels at the ordinary velocity of electric radiation (3×10^{10} cms./ sec.) and has, therefore, to be slowed down to a reasonable velocity which can be given to an electron stream without recourse to excessive accelerating voltage.

This is done by making the wave travel down a helix of wire so that although the wave travels round and round the helix turns at roughly its normal velocity, its axial velocity along the length of the helix is reduced to about one-tenth of that value. If the electron beam is sent down the centre of the helix (Fig. 59) it will be charge-density modulated, *i.e.* amplitude modulated, due to the interaction of the magnetic field of the helix with the electrons. This modulation grows along the length of the helix, roughly according to the square of the axial distance from the beginning of the helix. By suitable design this modulation can transfer energy to an external collector at the end of the helix and more energy can be extracted than was put in by the original signal, or in other words, the helix-electron beam mechanism is an amplifier. Some of the D.C. energy in the electron stream is being converted into A.C. energy in the wave.

In its practical form the valve consists of the wire helix supported in a long glass envelope which fits through two wave guide stubs for input and output. At one end of the tube is a conventional electron gun assembly, at the other a collector electrode. An accelerating voltage of about 2kV is usual with a beam current of 1mA. Power amplifications of up to 200 can be realised with the amazing bandwidth of 800 Mc/s. at an operating frequency of 4,000 Mc/s.

This very wide bandwidth is possible since the valve uses a completely untuned energy exchange system, in marked distinction from magnetrons and klystrons which use very sharply tuned resonant cavities. The bandwidth of a travelling wave valve is only limited by the coupling arrangements at the beginning and end of the helix. The valve opens up a new field of microwave endeavour, and, in particular, allows the use of R.F. amplification prior to crystal mixers in receivers. In television and multichannel microwave telephony it should prove invaluable. A single valve could handle, say, 10,000 telephone channels or about 30–40 high definition television programmes at the same time.

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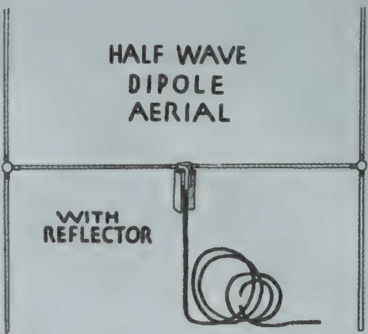

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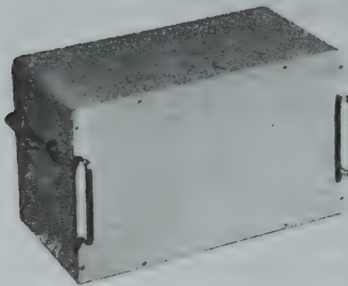

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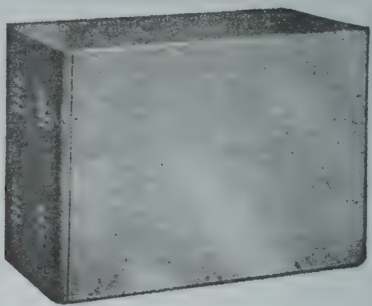

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VHF TECHNIQUE

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CONTENTS

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FOREWORD

VHF TECHNIQUE presents in convenient form a comprehensive account of the techniques which are employed for the generation, propagation, and reception of frequencies lying between 30 and 300 Mc/s.

For many years prior to 1940 radio amateurs had shown a great interest in such frequencies, and it is on record that their practical experience of operating conditions at that part of the spectrum proved of considerable value during the recent war.

The propagation characteristics of waves of very high frequency are now fairly well-known as the result of long-term amateur and professional investigation, but it would be unwise to assume that the full story has been written. As recently as the winter of 1947–48 radio amateurs succeeded for the first time in establishing two-way trans-Atlantic communications on frequencies of the order of 50 Mc/s. The importance of their achievement has been acknowledged by scientific bodies in Great Britain and the United States, who are well aware that the amateur fraternity, by reason of its numbers, enthusiasm and an urge to improve, is in the unique position of being able to make valuable contributions to existing knowledge. Scientific organisations are invariably limited in their scope because their observations are confined to a few widely-separated sites; amateur stations are found in almost every town and village.

Interest in VHF work is to some extent handicapped by the lack of suitable equipment, but, as the authors of this book have shown, it is frequently possible to obtain very satisfactory results with standard components and valves.

During 1948 and 1949 several new frequency bands will become available to radio amateurs including an allocation between 144 and 146 Mc/s. (2 Metres) This band will offer opportunities to the amateurs of the United Kingdom who have so far had no opportunity of exploring this part of the spectrum. The authors of *VHF Technique* have shown how this new band can be used to good effect.

The following is a list of the frequency allocations above 30 Mc/s. due to be released to British Isles amateurs:—

144–146 Mc/s. (2 Metres)	Exclusive.
420–460 Mc/s. (70 cm.)	Shared.
1,215–1,300 Mc/s. (25 cm.)	Exclusive.
2,300–2,400 Mc/s. (12·5 cm.)	Exclusive.
5,650–5,850 Mc/s. (50 mm.)	Exclusive.
10,000–10,500 Mc/s. (30 mm.)	Exclusive.

The band 2,300–2,450 Mc/s. is already available to British amateurs.

Readers who are interested in frequency bands higher than 300 Mc/s. are referred to *Microwave Technique*—a companion booklet in this series.

J. C.

CHAPTER 1 PROPAGATION OF VHF WAVES

IONOSPHERIC PROPAGATION—TROPOSPHERIC PROPAGATION —OTHER PHENOMENA.

RADIO waves of very high frequency (*i.e.* 50 to 500 Mc/s.) are subject to the same laws of propagation as other radio waves, but due to their higher frequency and correspondingly shorter wavelength the results achieved by their use often differ considerably from those obtained with waves of somewhat lower frequency. On the normal short-wave bands the dominating feature is ionospheric reflection and refraction, while on the VHF bands tropospheric propagation becomes of major importance.

Ionospheric Propagation

The ionosphere consists of two main ionised layers, the *E* layer at about 90 miles height, and the *F* layer at about 200 miles. During daylight the *F* layer usually divides into the *F*1 and *F*2 layers, and it is the *F*2 layer, at a height of up to about 300 miles, which is responsible for long distance short-wave communication over thousands of miles, by bending (or refracting) the waves sufficiently to return them to earth. As frequency is increased a greater degree of ionisation is required to effect the necessary degree of refraction, and a limiting frequency is reached above which waves are not bent sufficiently to be returned. This limit depends on time of day, season, direction and sunspot activity but is usually between 15 and 50 Mc/s. The upper limit is reached only at periods of high sunspot activity, and may on very rare occasions exceed 50 Mc/s. However, it is very unlikely that waves of frequencies exceeding 60 Mc/s. will ever be propagated around the earth's curvature by *F*2 layer ionisation.

The upper frequency limit for the *E* layer is normally much less than that for the *F* layer, but during the summer months in daylight hours a very intense ionisation known as "*Sporadic E*" occasionally occurs within the *E* layer. This ionisation is capable of reflecting frequencies up to about 100 Mc/s. and enables communication to be effected over distances from about 400 to 1,200 miles. On lower frequencies, *e.g.* 14 Mc/s. the "skip distance" may disappear altogether. Signals received by this means are frequently very strong, but may be subject to severe fading. The appearance and disappearance of "*Sporadic E*" is quite erratic and unpredictable. Further, it may be local in occurrence. It is thus not able to produce reliable channels of communication. When using rotary beam aerials best results are not always obtained by pointing the beam in the great circle direction of the distant station.

During intense ionosphere storms, such as are accompanied by auroral displays, signals of the order of 30 to 60 Mc/s. have been received by reflection from some northerly point. Such signals are characterised by a noise modulation, perhaps best described as a "burble." To enable contact to be made by this means both transmitter and receiver aerials must be directed to the same northerly direction if beams are employed. It is presumed that these reflections are from the stream of ionised corpuscles entering the earth's atmosphere after their journey from the sun. The upper frequency limit for this phenomenon is uncertain, and there is scope for investigation.

Tropospheric Propagation

The troposphere is that part of the atmosphere where weather is the predominating feature. As height is increased the density of the atmosphere

slowly decreases, and in consequence the refractive index of the air also decreases. This means that the paths of rays travelling through the atmosphere will be slightly curved, the curvature being towards the earth. This curvature results in the actual horizon being more distant than the geometrical horizon. This is shown in Fig. 1. The actual increase in distance is normally about one-seventh. Beyond this extended horizon, signals can be received by diffraction—an effect which decreases in intensity as the frequency is increased. The range of the extended horizon depends upon (*a*) the height of the transmitting aerial, (*b*) local topography, (*c*) obstructions (all of which are probably stable), (*d*) the temperature, and (*e*) the humidity characteristics of the layer of air near the earth's surface. These last two factors are variable from day to day, and even minute to minute, and affect the value of the refractive index of the air. Hence, there may be noticeable changes in the extended horizon distance which will cause appreciable fluctuations in field strength at any given point in its neighbourhood. The normal extended horizon

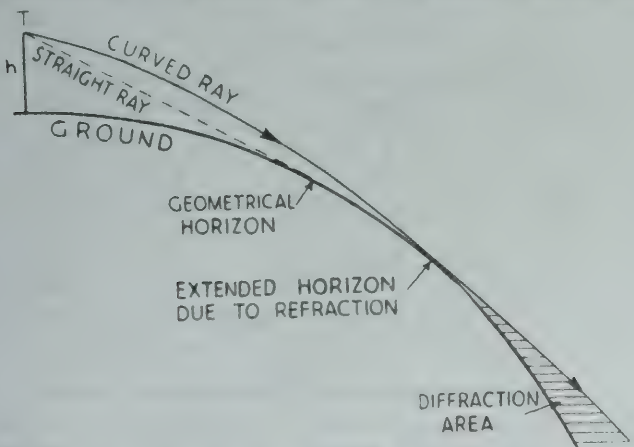


Fig. 1.
Diagram showing the extension of horizon distance due to refraction in the lower atmosphere.

distance, neglecting obstructions, is given by the formula $1 \cdot 42\sqrt{h}$, where h is the height in feet of the transmitting point, the horizon distance being in miles. If both transmitter and receiver are elevated, the extended horizon distances should be added to ascertain whether they are within " line of sight." Due to the greater height of the aerial compared with wavelength, the field strength at the horizon may increase when frequency is increased, an effect which will be considered in more detail when dealing with aerials. Beyond the horizon, however, the field strength will fall off more rapidly on higher frequencies.

Transmission to places beyond the extended horizon and diffraction area is possible when certain types of abnormal humidity and temperature gradients occur in the troposphere, below about 5,000 feet. The normal rate of decrease of refractive index in the atmosphere with height is of the order of 12×10^{-6} per 1,000 feet. The decrease required to make a horizontally radiated ray curve sufficiently to follow the earth's curvature

is 48×10^{-6} per 1,000 feet. The rate of decrease is more rapid when the water vapour content of the air decreases rapidly, that is, when a layer of dry air overlies a moist layer. A temperature inversion accompanying such a humidity gradient may assist, but an inversion without a humidity gradient can have an adverse effect.

Fig. 2 shows the effect of humidity and temperature changes on the refractive index of the air during the night of October 11/12, 1946. It will be seen that from about 1,800 feet to 2,600 feet the rate of fall of the refractive index averages 44×10^{-6} per 1,000 feet. On that particular night signals were received on 60 Mc/s. at distances up to 250 miles. It is probable that these signals were returned to earth by partial reflection in the layer, the increased refraction assisting a condition of grazing incidence (*i.e.* only a fraction of a degree from horizontal), an essential for adequate reflection at a diffuse boundary. Since the incident radiation in this case is that which left the transmitting aerial in a very nearly horizontal direction, the vertical polar

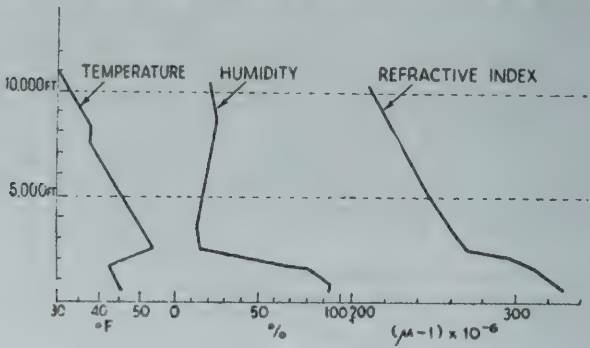


Fig. 2.
Tropospheric characteristics for midnight October 11/12, 1946.

diagram of the aerial must be such as to provide as much near-horizontal radiation as possible if full advantage is to be taken of these conditions. This will be considered further in the chapter describing aerial systems.

With increased intensity of the incident radiation some reflection is obtainable with refractive index gradients much less than that shown for October 11/12, and in consequence some stations operating on 60 Mc/s. have found it possible to maintain almost unbroken schedules at distances well beyond the normal diffraction area. Field strengths are, however, liable to large day-to-day variations. Due to the shorter wavelength the intensity of such reflections will be less on higher frequencies, but at these higher frequencies there may form around the layer a duct of sufficient width to act as a wave guide, in which the very short waves are trapped and travel round the earth's curvature. Fig. 3 shows this on an exaggerated scale. From a height $H1$ up to $H2$ above the earth the rate of decrease of the refractive index exceeds 48×10^{-6} per 1,000 feet and can, therefore, produce a bending of rays greater than the curvature of the earth. The effect of this on three rays is shown. Ray 1 enters the layer at too large a grazing angle and passes through the layer without reaching a horizontal path. Ray 2, at a smaller grazing angle, is bent just to horizontal at the top of the layer and will, therefore,

continue to follow the earth's curvature. Ray 3 at a still smaller angle reaches horizontal condition at point A where the rate of decrease of refractive index is near its maximum and it continues to increase its curvature and bends towards the earth. From point B the rate of curvature of the ray falls off considerably and, with a suitable gradient of refractive index, point C, where the ray is once more horizontal, may be reached before the ray reaches the ground. It will then be bent upwards again and continue its oscillatory path around the earth. The effect is greatly enhanced when the transmitter is actually in the duct, *i.e.* when the duct is at ground level, or the transmitter elevated.

Investigations made in southern England during 1946 showed that gradients of refractive index in excess of 48×10^{-6} per 1,000 feet are extremely rare, and thus it is considered that the " partial reflection " theory offers the better explanation of the fairly regular reception of 60 Mc/s. signals at distances up to 300 miles.

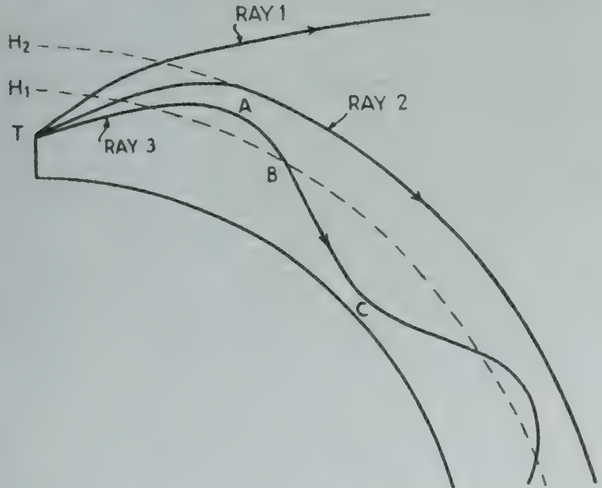


Fig. 3.
An exaggerated illustration of ducting.

Other Phenomena

Rapid and severe fading is occasionally experienced on VHF signals —including even local signals—due to reflection from aircraft. As the aircraft moves, the reflected signal changes its phase with respect to the directly received signal due to the changing path-difference. The periodicity of this fading depends upon the speed of the aircraft, its course and the frequency of the transmission involved, but periodicities of the order of one second are not uncommon.

Another phenomenon often encountered when listening to weak transmissions is sudden short bursts of signal (usually, only for a fraction of a second) at great strength. This is sometimes known as the " shooting-star " effect, because it is thought to be due to reflections from the trail of ionisation

left behind a meteor as it passes through the ionosphere. During big displays of shooting stars the ionisation may become sufficiently continuous to permit telephony communication to take place over distances up to about 1,000 miles, although c.w. transmissions may be too mutilated to be readable.

Summary

Reception of signals of frequencies between 60 and 100 Mc/s. and possibly higher, at distances up to 1,200 miles is occasionally possible due to ionospheric abnormalities, while the normal ground wave range is frequently extended by several hundred miles by tropospheric phenomena. This extension applies throughout the VHF band although the exact mechanism may differ between 60 and 500 Mc/s.

CHAPTER 2 VHF AERIAL SYSTEMS

LOW ANGLE RADIATION—POLARISATION—SIMPLE AERIALS—BEAM AERIALS—FEEDING PARASITIC BEAM AERIALS—FEEDER LOSSES—ADJUSTING THE BEAM—COUPLING TO THE TRANSMITTER—OTHER AERIALS.

THE principles underlying the design of aerial systems for use on very high frequencies are in no way different from those which apply to somewhat lower frequencies, therefore types of aerial used on such bands (viz. 7 to 30 Mc/s.) can also be employed at VHF. However, due to the smaller physical dimensions involved, certain types which are impracticable on lower frequencies become reasonably easy to build and erect for use in the VHF bands, and it is with this latter type that this chapter will be mainly concerned.

Low Angle Radiation

A perusal of the previous chapter, where the propagation of VHF waves was considered, shows that it is desirable to radiate as much energy as possible at extremely low angles. For tropospheric propagation, radiation at angles above 1° from horizontal is wasted power. On the high frequency bands where ionospheric transmission occurs, the desired radiation angles are from about 5° to 15°. This, of course, applies to "*Sporadic E*," but most VHF aerials will radiate at these angles without any trouble, and it is not usual to design aerials especially for that type of propagation.

The intensity of very low angle radiation is effectively increased by elevating the transmitter aerial. Fig. 4 shows the vertical polar diagrams for a horizontal half-wave aerial at various heights above a surface of perfect conductivity—sea water being the nearest approach to such a surface actually encountered. It will be noted how the vertical angle of the lowest lobe decreases as the aerial is raised. These lobes are due to reflection from the ground and Fig. 5 enables the derivation of these diagrams to be understood. It will be seen that the path difference between the direct and reflected waves is equal to $2H \sin A$, where A is the angle above horizontal and H is the height of the aerial. For ground of perfect conductivity and horizontal polarisation there is a phase-reversal at reflection, so that for the direct and reflected waves to reinforce each other $2H \sin A$ must be equal to an odd number of half wavelengths, or for the bottom lobe,

$$\sin A = \frac{\lambda}{4H}$$

both λ and H being in the same units. For example, to make A as small as 1°, using a frequency of 60 Mc/s., the aerial needs to be more than 200 feet high. This is, of course, the major point of the lobe and there is radiation at angles below this point. For ground other than that of perfect conductivity the polar diagram is somewhat distorted and the minima may not drop to zero due to the coefficient of reflection departing from unity, but the same general principle holds.

The desired increase in height may be obtained by careful choice of transmitter site. In this respect high ground sloping away rapidly, in order to avoid ground reflections in antiphase from levels comparable with transmitter height, appears to be particularly suitable. Such has been borne out

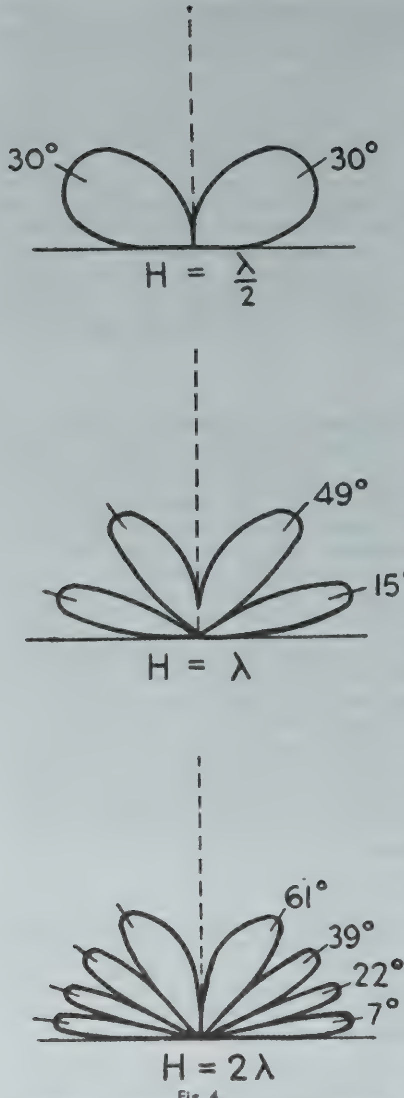


Fig. 4.
Vertical polar diagrams of horizontal half-wave aerial at various heights above ground of perfect conductivity. Aerial at right angles to plane of paper.

by the success of amateur stations working on 60 Mc/s. from such locations. Most amateurs will not be in a position to operate from such sites except during field days, in which case the aerial should be erected as high as possible and clear of all ground obstructions. The trouble involved in doing this is well repaid, and the principle applies to both transmitting and receiving aerials. Fig. 6 shows the aerial in use at one of the most successful 60 Mc/s. stations in southern England and gives a good illustration of the desirable position for a VHF beam aerial.

The above remarks should not completely discourage amateurs who are unable to erect such high aerials from using the VHF bands, for useful results have been obtained with beams in attics and on low garden masts, but stations using low aerials must not expect to be able to work consistently much beyond horizon range. Low aerials should only be used when a higher aerial is an impossibility.

Polarisation

The vertical polar diagrams depicted in Fig. 4 were for horizontal aerials and it may be thought that the desirable horizontal radiation might be more readily obtained by using vertical aerials. Much will depend on the nature of the ground particularly its electrical properties (*e.g.* dielectric constant and conductivity, which vary with different soils) and its smoothness which affects the reflection coefficient, so that an accurate prediction is difficult to make. With vertical polarisation and small grazing angles at VHF there is considerable phase-change at reflection, while the reflection-coefficient decreases rapidly as the angle increases. Hence,

although at lower frequencies vertical polarisation may give greater intensity of radiation at near-horizontal angles, at VHF there is often little to choose between vertical and horizontal, while sometimes horizontal polarisation may be superior.

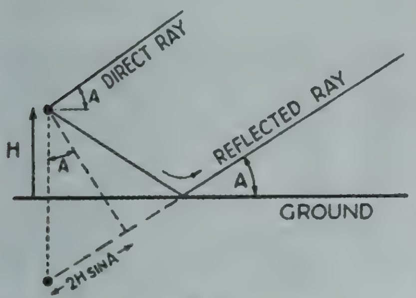


Fig. 5.
Reflection from ground. Path difference between direct and reflected wave is seen to be $2H \sin A$.

Some tests made from the Empire State Building, New York, using frequencies of 49·5, 83·5 and 142 Mc/s., over land and including both town and country areas showed that horizontal polarisation (1) varied over a greater range, (2) had higher maximum values, (3) produced a higher average field strength than vertical, other things being equal. Results over a sea path might, of course, produce different conclusions.

In addition, vertical polarisation is subject to more interference from man-made static. Considerable fluctuations in signal strength over short distances may be expected in towns—particularly when the aerial is low—due to reflections from buildings, overhead wires, etc. Both polarisations are liable to trouble from this source. Finally, whichever type of polarisation is used, it must be the same at both transmitting and receiving stations. On 60 Mc/s. most British amateur stations employ horizontal polarisation.

Simple Aerials

Half-wave and long wire aerials may be used for VHF work with similar methods of feed to those used on higher frequencies. The length of a half-wave can be found from the formula :—

$$\textit{Length (in inches)} = \frac{5540}{F},$$

where F is the frequency in Mc/s. On the higher frequencies it may be necessary slightly to reduce the figure obtained by this formula, while a long wire containing n half-waves will be rather longer than n times the figure given by the formula. The discrepancy obviously increases with the number of half-waves and the approximate length required is shown in Table I.

Fig. 6.
The 3-element rotary beam used at G5MA Ashstead, Surrey, on 60 Mc/s.

Tuned feeders are not generally advised, but if they are used the spacing must be reduced from that employed on lower frequencies. Six inches is no longer

a negligible fraction of a wavelength at 100 Mc/s. and with feeder spacing of that order there will be radiation from the feeders.

The directional properties of aerials are usually found to be very marked on these frequencies. This is due to the reliance on very low angle radiation,

TABLE I.

Properties of Long Wire Radiators.

No. of Half-waves.	Length in feet (f in Mc/s.).	Angle of Major Lobe.	Angles of Zero Radiation.	Gain in Major Lobe over Half-wave (db).
1	462/f	90°	—	—
2	954/f	54°	90°	0·6
3	1,446/f	43°	70°	1·1
4	1,940/f	36°	60°, 90°	1·6
6	2,920/f	30°	49°, 71°, 90	2·6
8	3,906/f	26°	41°, 60°, 75°, 90°	3·5
10	4,890/f	22·5°	37°, 53°, 66°, 78°, 90°	4·3

The angles are measured from the direction of the wire. In addition to the angles of zero radiation given above, in all cases there is a minimum of radiation from the ends (*i.e.* at 0°).

so that the high angle and end-fire radiation, often useful in ionospheric propagation, is non-effective. For this reason it is desirable to make half-wave horizontal aerials rotatable, and care must be taken to see that the major lobes of long-wire aerials are positioned in useful directions and nulls do not fall on centres of activity. Long-wire aerials cannot be considered as omni-directional, and can, with more accuracy, be described as fixed beams, with appreciable gain in the major lobes over a half-wave. The angles of the major lobes from the line of the aerial are given in Table I together with the directions of zero radiation and other useful information, while some examples of horizontal polar diagrams for long wires are given in Fig. 7. It should be noted that the major lobe is always that nearest the direction of the aerial.

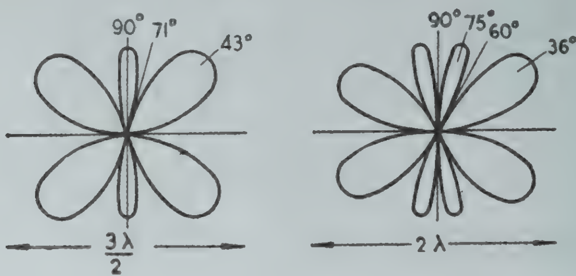


Fig. 7.
Polar diagrams of some long wire aerials.

Beam Aerials

Although the simple aerials mentioned above will give results on the VHF bands, most amateurs working on these frequencies will wish to take advantage of the more efficient aerial systems made possible by the shorter wavelengths in use. On the 60 Mc/s. band it is probably no exaggeration to say that 80% of the stations regularly active use a rotary beam. Usually the beam is one making use of half-wave parasitic elements. A parasitic element is one which is not energised directly from the transmitter, but which receives its power by induction or radiation from a neighbouring element, usually known as the driven element. This driven element naturally derives its power from the transmitter usually *via* some type of feeder. Such beams are easy to construct and no great difficulty should be encountered in adjusting them for satisfactory operation. They give worthwhile gains (up to about 9 db for 3 elements), particularly at low horizontal angles.

The rotating mechanism of the beam array shown in Fig. 6.

Parasitic elements can be arranged to act as directors or reflectors. An element forward of the driven element in the direction of maximum radiation is called a director, while an element to the rear is a reflector.

Considering first the requirements for a reflector it might appear that the correct spacing between the driven element and the reflector would be a quarter wavelength, allowing for the time of travel of the radiation between the elements and the phase reversal at the reflection. There are two main reasons why this spacing is not the correct one for maximum gain in the forward direction. Firstly, the voltage induced in the reflector is due not only to radiation from the driven element, but also to the electric and magnetic fields surrounding it. This has serious effects on the phase of the induced e.m.f. and current. Secondly, moving the reflector closer to the driven element increases the current induced, which more than compensates for the incorrect phasing thereby

General view of the mast and rotating mechanism of the beam array shown in Fig. 6.

produced, so that the radiation in the desired direction is increased.

The optimum spacing is found in practice to be between ·15 and ·25λ. The phase of the reflected radiation will be slightly leading on the radiation from the driven element due to the close spacing. This can be corrected to some extent by mistuning the reflector so that its reactance is inductive, *i.e.* by lengthening it. An increase in length of about 5% is usually found to be about correct, with the spacing at ·15λ.

If the parasitic element is brought closer than about ·14λ then radiation is reinforced in the opposite direction and the element becomes a director. This effect is enhanced if the length of the director is shortened by about 4%. A spacing of ·1λ is commonly used.

Both director and reflector may be used at the same time, and more than one director is not uncommon. As additional elements are added the beam becomes sharper so that small errors in directivity cause a large reduction in signal strength, although at the same time the gain in "the line of shoot" is increased. Some measurements made on a 3 element close-spaced beam on 60 Mc/s. showed the field strength to be 6 db down at appoximately 30° either side of the line of shoot. At angles greater than this the strength fell rapidly, being more than 30 db down at right angles to the forward direction.

TABLE II.

Gains to be expected from simple beams.

No. of Elements.	Director Spacings.	Reflector Spacings.	Possible Gain (db).
3	·1	·15	7 to 8
3	·2	·15	8 to 9
3	·2	·2	9
4	·1	·15	8 to 9
4	·2	·2	9 to 10

For beams employing more than two elements, the optimum spacings given above require modification. The gains to be expected from a number of different spacings of 3 and 4 element beams, adjusted for maximum performance are given in Table II. From this it will be seen that although useful gain can be obtained with close spacing of elements, for best results, spacings of the order of ·2λ are advisable.

When constructing a beam, it should be arranged so that some slight adjustment (say 5%) of element length is possible after erection. In this way maximum performance can be obtained, and it is recommended that these adjustments be made by observing the effects on a field-strength meter placed as far as possible from the beam. The reading on the meter can be observed through field glasses, or by bringing back the actual meter to the neighbourhood of the beam by means of flex. In this latter case precautions must be taken to ensure that no radiation is being picked up by the flex, as this will give misleading results.

It should be noted that, due to the compromises between correct phasing and maximum induced currents, the position and lengths of the elements will not generally be found to give maximum forward gain and maximum front-to-back ratio at the same time. For transmission purposes maximum gain is the usual requirement, and this adjustment will be found to be accompanied by a very reasonable front-to-back ratio. More detailed notes on the adjustment of element lengths, etc., will be given later, after methods of feeding multi-element beams have been considered.

For reliable operation, beams should be constructed from material that will not bend excessively in high winds. It is important that the spacings remain constant, otherwise erratic results will be obtained. On 60 Mc/s. ½" tubing

is recommended, while on higher frequencies some reduction in diameter is permissible from the rigidity point of view, as the total length of tubing becomes less. Duralumin tube is very suitable, being both light and a good conductor, but it is liable to severe corrosion and a protective coating of aluminium paint may be desirable. Before paint is applied to duralumin the surface must be treated in such a way as to provide a key to the paint. This can be done by roughing in a sand blast or with a scratch brush. Further details on this subject may be obtained on application to the *Aluminium Development Association*, 67, Brook Street, London, W.1. Care must be taken with joints to obtain as much contact area as possible because, due to the presence of a strongly adherent oxide film, such joints are liable to show a high resistance. The three elements of the beam may be held together by a wooden frame-work, suitably treated to withstand the weather, the elements being mounted on stand-off insulators. Alternatively, the centres of the elements can be joined together rigidly by a length of wide duralumin tubing, needing no insulators. This is possible since the centres of the elements are all at the same RF potential.

Fig. 8.

Input resistance of driven element for various lengths of director at two different spacings in a four element beam Curve A was with directors spaced 0·2λ reflector 0·15λ. Curve B for directors 0·1λ reflector 0·15λ. Note.—This diagram should not be used to design a beam as the input resistance depends also on other factors.

Feeding Parasitic Beam Aerials

The problem of exciting the multi-element beam, particularly when close spacing is in use, is complicated by the extremely low centre-impedance of the driven element. The exact value of this impedance depends on numerous factors, including the length and spacing of the directors and reflectors. Fig. 8 shows some figures obtained experimentally with a 4 element beam. This diagram is not intended to provide data for designing a beam, but to emphasise the large variations which may occur when spacings or lengths of elements are altered. The low impedance not only makes direct feed difficult but means that the beam will only be effective over a narrow band of frequencies. Delta matching can be used, and has been used effectively on 60 Mc/s., the tapping points being found by trial and error, but the excessive fanning-out may result in appreciable radiation and so have a detrimental effect on the polar diagram of the beam.

In order to facilitate feeding, the most general method is to use a folded dipole. In addition to making possible feed from the more usual types of high and low impedance line, the folded dipole method broadens the frequency band over which the aerial can be used. In this system a number of radiators

are closely spaced (usually about 5 times their diameter) and arranged so that they contribute in-phase components of radiation. Fig. 9 shows a folded dipole consisting of two wires. The radiation resistance at the feed point increases in proportion to the square of the number of wires used,

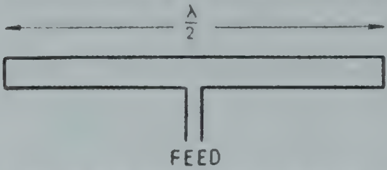
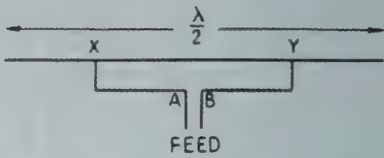


Fig. 9.
The "two" wire folded dipole.

provided they are all the same diameter, and hence with a two wire system it is four times that of a single wire. Thus by employing a folded dipole as the driven element of a beam a more reasonable figure of radiation resistance is obtainable. The figure is also affected by the wire or tube-spacing, particularly if this is too small, while a further variation is possible by the use of different diameter conductors for the two sections of the folded dipole. By making the broken (or "fed") conductor narrower than the unbroken one the impedance multiplying factor is increased. Provided the two con-

Fig. 10.
The "T" match aerial.



ductors are not too close the approximate multiplying factor can be found from the formula

$$\left(1 + \frac{Z_1}{Z_2}\right)^2$$

where Z_1 is the characteristic impedance of a line formed by two conductors both like element 1 and with same centre-to-centre spacing as elements 1 and 2, and Z_2 is the characteristic impedance of a line similarly formed by conductors like element 2. In this, element 1 is the "fed" element. Thus with the unbroken element made of $\frac{1}{2}''$ tube and the fed element 12 S.W.G. wire the multiplying factor becomes 7·8. Using a narrower gauge wire will produce a greater multiplying factor. By choice of a suitable combination of conductors it is, therefore, possible to obtain almost any desired input resistance.

A second method of feeding a close spaced beam aerial, which is worthy of mention is by "T match". This is shown in Fig. 10 which will be seen to be somewhat similar to the folded dipole already considered. High impedance feeders may be used, and the position of the feed points X and Y obtained experimentally so that standing waves on the feeder are eliminated. The spacing may be about $1\frac{1}{2}''$ to $2''$ at 60 Mc/s. and proportionately less at higher frequencies. The matching sections AX and BY can be either heavy gauge wire or tubing similar to the rest of the beam. The gauge of this conductor will affect the positions of X and Y, but XY will be of the order of $\frac{1}{8}\lambda$ when using a 300 ohm line.

QUARTER WAVE MATCHING SECTION

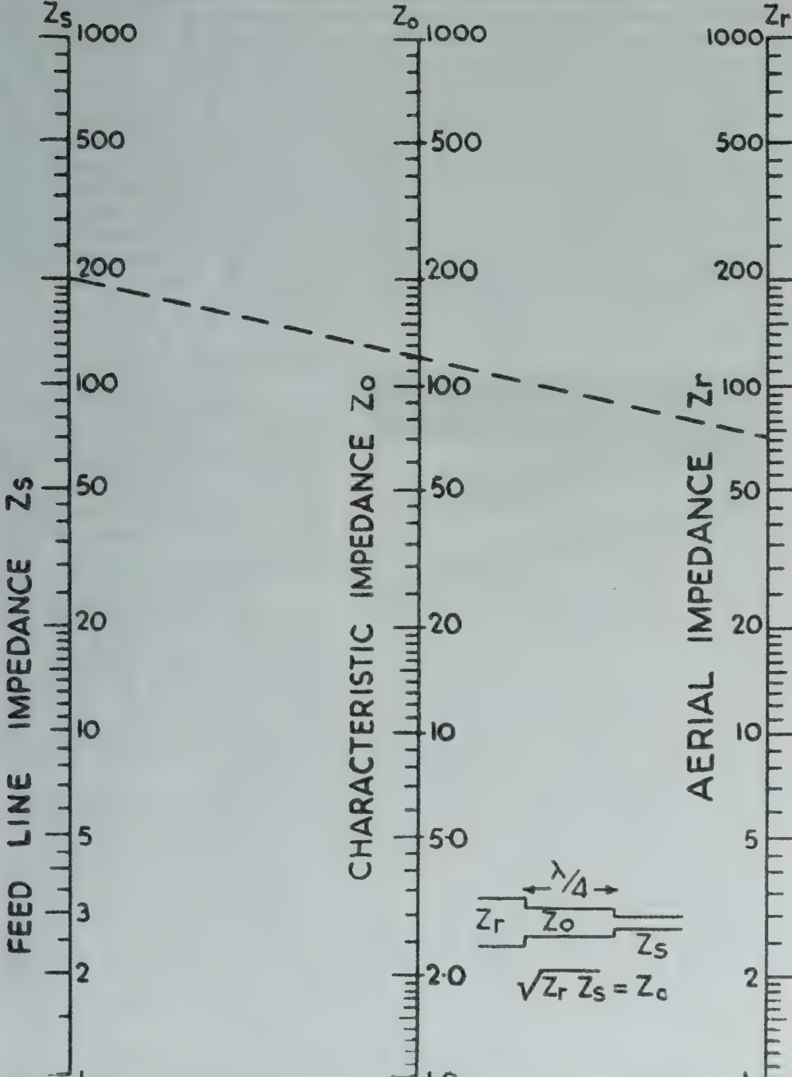


Fig. 11.

Matching section chart. This chart may be used to obtain the surge impedance of a quarter-wave matching section used as an impedance transformer from one real impedance to another. In the example shown, *Zr* is 72 ohms and *Zs* is 200 ohms, indicating a quarter-wave matching section of 120 ohms is needed.

Direct feed to the centre of the beam is possible by using a quarter-wave matching transformer, which can conveniently consist of an electrical quarter wavelength of low impedance feeder. Such a length of 75 ohm cable will correctly match a 600 ohm line into 8·5 ohms and will give a standing wave ratio of less than 2 to 1 over a range of from 5 to 16 ohms. A chart, enabling correct combinations of cable for the matching transformer and for the feed line to be employed, is given in Fig. 11, but as the exact impedance at the input to the beam is a matter of some doubt, the final correct combination must almost certainly be found by trial and error. It should be noted that an electrical quarter-wave of cable is less than a physical quarter-wave due to the slower speed of travel of waves in the cables than in air. The physical length, corresponding to a required electrical length, can be obtained from manufacturer's data or ascertained experimentally as described below. With 75 ohm cable, using high quality dielectric, an electrical wavelength is usually about 60 to 70% of a physical wavelength.

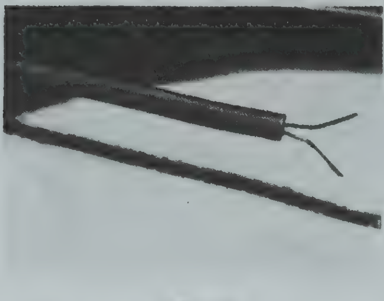


Fig. 12.
A balanced 80 ohm feeder cable. Loss at 56 Mc/s. 4 db per 100 foot run. (*Belling and Lee, Ltd.*)

Feeder Losses

In choosing cables for use on very high frequencies it must be remembered that, mainly as a result of the increasing "skin effect," transmission losses increase as frequency is increased, and cable which has negligible loss on, say, 7 or 14 Mc/s. may show up very badly at 100 Mc/s. Losses of 5 db per 100 feet are common at this frequency in cables of quite good quality, *e.g.* the 80 ohm line shown in Fig. 12. This is, of course, the loss under conditions of perfect matching to the aerial, but if the matching is incorrect the loss may increase very considerably and in very bad cases very little power indeed will reach the aerial itself.

Fig. 13 shows, in graphical form, the effect of standing waves of various magnitudes on the total feeder losses, the loss under matched condition being that given at the extreme left of each curve. For example, curve A is for a length of cable having a total loss under perfect matching conditions of 1 db. With a standing wave ratio of 4 to 1 the total loss increases to 2·5 db (probably not very serious), but if the matched line has a loss of 2 db then a similar standing-wave ratio increases the total losses to 7 db. Hence it will be seen that standing-waves become a serious matter if the cable length is large. Thus, it is useless to expect to gain in field strength at a distant point by increasing the height of the aerial if losses are thereby incurred in the feeder.

With a low aerial, using 30 feet of cable correctly matched, the feeder loss will be of the order of 1·5 db. If there is a mismatch of 2 to 1 this loss increases to 2·5 db. If the same aerial is raised to 100 feet then with correct matching the total loss is now 5 db, while with 2 to 1 mismatch the loss is as large as 8 db. This must be offset against any radiation gain due to the extra height.

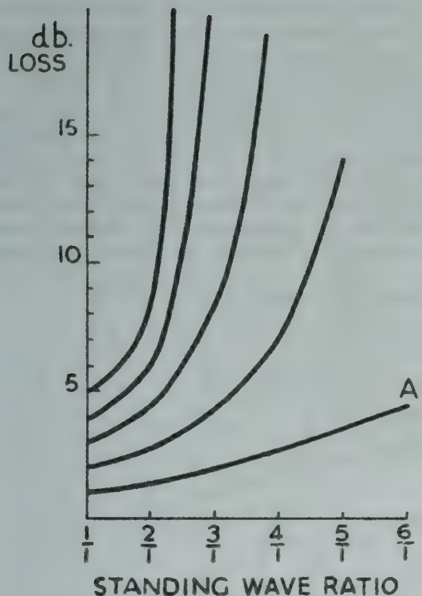


Fig. 13.
Increase of loss with standing waves on feeders. See text.

Adjusting the Beam

Before attempting to adjust the parasitic element lengths it is essential to ensure that the driven element is resonant and also fairly closely matched to the feeders. As all director and reflector adjustments are liable to affect both the resonance and impedance of the driven element it is also necessary to be able to make frequent checks. It is suggested that the simple circuit shown in Fig. 14 be constructed and coupled to the transmitter power amplifier. The points X and Y should be connected by a non-inductive resistor of approximately the same value as the feeder surge impedance. Due to skin effect the actual resistance to r.f. currents will not be exactly the same as that given on the resistor. The circuit should be tuned to resonance as indicated by the maximum reading on the ammeter, the coupling being adjusted to a suitable value for the wattage of the resistor in use. It is now necessary to connect a length of feeder an exact number of electrical half-waves long to points X and Y, in place of the resistor. This is done by connecting a length rather longer than that anticipated and cutting off pieces until the tuning of the circuit for resonance is exactly the same as with the resistor. (The reading on the meter may be slightly different due to skin effect in the resistor.) If only one half-wave is required the starting length may be about 80% of a physical half-wave. Having obtained a feeder a number of half-waves long, connect the aerial to the far end, when, if it is resonant the tuning of the coupling circuit should remain unchanged. To check for matching, lengthen the feeder by a quarter wavelength. With this done the tuning and meter reading should remain the same.

The feeder should now be connected to the transmitter in the normal way, a field strength meter placed at a suitable distance and the director and reflector adjusted for maximum reading. It will probably be found that the length of the director is much

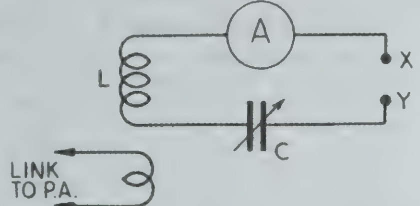


Fig. 14.
Simple device for checking resonance of aerial and accuracy of matching of feeders to aerial. L and C should be of suitable values to be resonant at frequency in use. For method see text.

more critical than the reflector. While doing these tests it is necessary to make frequent checks to see that the driven element is still in resonance and correctly matched otherwise misleading results may be obtained. With more than one director it may be found that making the forward director a little longer than the others will give improved results.

Fig. 15 shows the effect of increasing and decreasing the director length from a basic length of ·48λ—the driven element being maintained resonant. These curves are for two different spacings and 4 element beams similar to those considered in Fig. 8. Although some slight extra gain may be obtainable with certain spacings by making the director longer than the usual shortened length, the sudden decrease in performance at slightly longer lengths will restrict the band of frequencies over which the beam can be used.

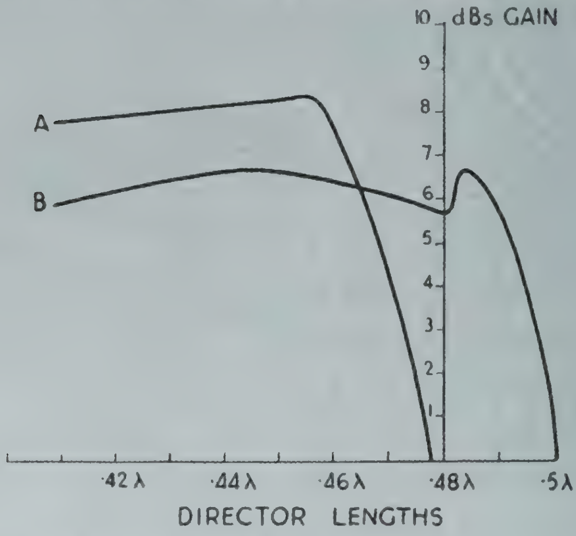


Fig. 15.
Actual gains obtained with two different spacings in four element beam arrays, with various director lengths. For details of array see Fig. 8.

Coupling to the Transmitter P.A. Circuit

When coupling the transmission line to the output circuit of the transmitter, care must be taken to avoid undesirable capacity coupling. This can cause the radiation of strong harmonics and seriously upset the working of the aerial system on the fundamental frequency. Due to the fact that capacitive reactance decreases when frequency increases these effects assume larger proportions on the VHF ranges than on lower frequencies.

The most usual method of coupling the feeder to the transmitter output circuit is by means of a one or two turn coupling coil. The voltages produced at the two ends of this coil are in antiphase and hence current in antiphase flows in the two feeder lines and no feeder radiation occurs. When stray capacitative coupling is present it produces voltages of substantially the same phase throughout the coupling coil and the two feeders are fed in

parallel, *i.e.* current in phase on the two lines and therefore, radiation from the feeder occurs. This radiation will become large if the length of the feeder, plus aerial top, is resonant at the frequency in use and the polar diagram of the aerial system will be altered. This means wasted power. Capacitative couplings should be suspected if, in spite of all attempts to match the feeders to the aerial, the line currents remain unbalanced and large standing waves persist.

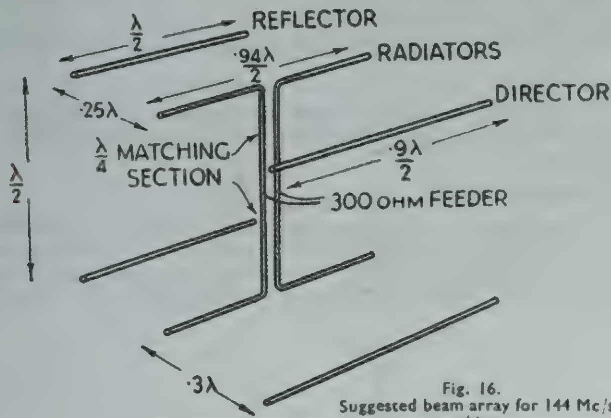


Fig. 16.
Suggested beam array for 144 Mc/s. working.

Harmonic radiation is also frequently due to this cause, the feeder length being resonant at the harmonic frequency. Since harmonics from amateur frequencies in the VHF region almost always fall in bands allotted to other services, the desirability of eliminating them is obvious. It must be stressed here that the use of push-pull output stages is no guarantee that even harmonics cannot get into the aerial system.

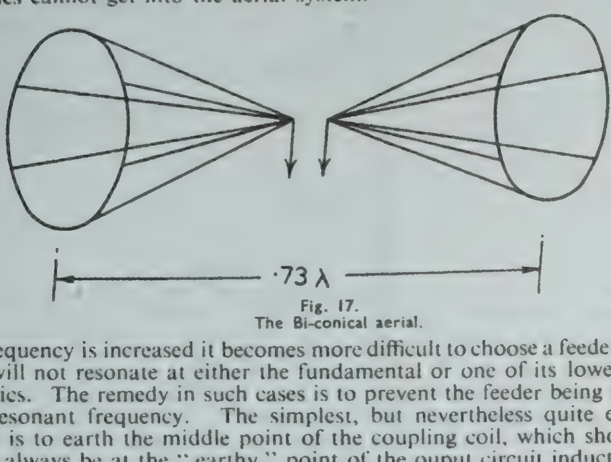


Fig. 17.
The Bi-conical aerial.

As frequency is increased it becomes more difficult to choose a feeder length which will not resonate at either the fundamental or one of its lower value harmonics. The remedy in such cases is to prevent the feeder being excited, at its resonant frequency. The simplest, but nevertheless quite effective method is to earth the middle point of the coupling coil, which should, of course, always be at the " earthy " point of the ouput circuit inductance.

Other Aerial Systems

The systems described above are those most commonly used by amateurs in the 60 Mc/s. band, but many other arrangements can be used with good effect. Parasitic type beams of a more complex type can be built by stacking arrays one above the other, half a wavelength apart, and fed in phase. This gives a noticeable increase in low angle radiation. A suggestion for such an array for 144 Mc/s. is given in Fig. 16. This array should work equally well either horizontally or vertically. Construction can be of ¼" tubing and the matching can be adjusted by varying the spacing of the ¼λ matching sections.

Bi-conical Aerial

Fig. 17, which illustrates an aerial suitable for wide-band operation, consists of two multiple wire cones and has an input impedance of about 60 to 70 ohms.

Co-axial Aerial

An efficient aerial with an omnidirectional characteristic and low angle radiation, suitable for use when vertical polarisation is desired is shown in Fig. 18. The centre conductor of a 72-ohm co-axial cable is extended for ¼λ to act as the upper half of a half-wave vertical aerial. The lower half is

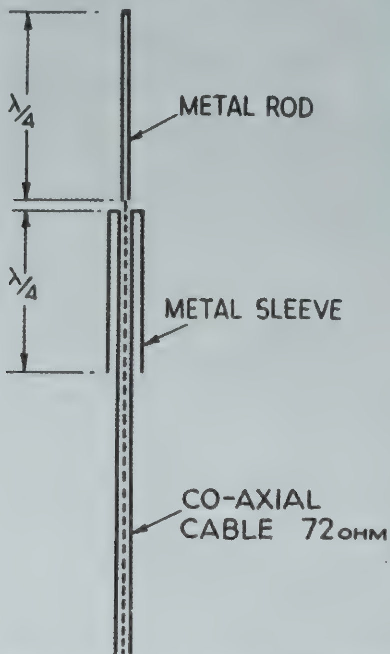


Fig. 18.
The coaxial aerial. This aerial produces good low angle radiation with vertical polarisation.

obtained by arranging a metal sleeving around the top ¼λ of the cable with the sleeve connected to the top of the outer conductor of the cable. There is thus a perfect match from feeder to aerial.

Corner Reflectors

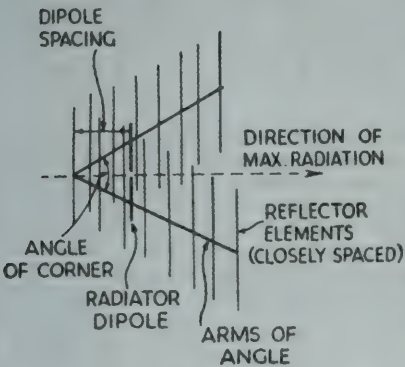


Fig. 19.
The corner reflector aerial.

The corner reflector consists of two flat conducting sheets which intersect at an angle. The sheets can be continuous metal or made of a number of parallel wires, which need not be electrically connected. The radiator is placed on the bisector of the "corner," and this bisector will be the direction of maximum radiation. The distance between the ½λ radiator and the vertex for a 60° corner angle should be a half wavelength, and under this condition the input impedance will be 70 ohms and a gain of about 10 decibels has been obtained in practice at 250 Mc/s. The cover illustration shows a 250 Mc/s. 60° corner reflector aerial. The length of each arm of the angle should preferably be about three times the spacing of the dipole from the apex; and the "herringbone" reflector elements should be longer than a half wavelength. Various angles and dipole spacings may be used and further information may be obtained from the original paper by Kraus in *Proc. I.R.E.*, Vol. 28, p. 513, November, 1940, or from Terman's *Radio Engineer's Handbook*, pp. 819–821. The salient features of the corner reflector are shown in Fig. 19. Practically, the corner reflector is feasible at frequencies above about 100 Mc/s. Below this frequency it is rather large. It may be constructed on an "A" shaped wooden frame, details of which can be gleaned from the cover illustration. Fig 20 shows a suitable balanced-to-unbalanced matching device which may be used to feed a dipole such as that used in a corner reflector, from an unbalanced co-axial line. Some distortion of the theoretical polar diagram must be expected if a dipole is fed from a co-axial line without the special matching arrangement.

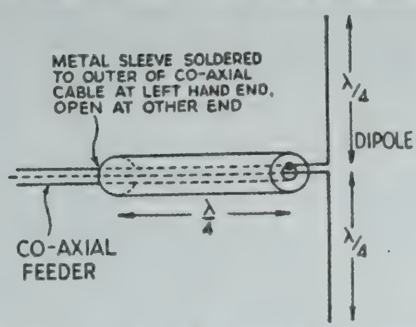


Fig. 20.
A simple balanced-to-unbalanced matching transformer.

CHAPTER 3 VHF TRANSMITTERS

INTRODUCTION — BY-PASSING — VALVES — NEUTRALISING — FREQUENCY MULTIPLIERS—TUNED CIRCUITS—PRACTICAL TRANSMITTERS — MODULATION — BCL INTERFERENCE—420 Mc/s.

Introduction

THE appearance during recent years of a number of valves especially designed for use in low power VHF transmitters has greatly simplified the design and construction of such equipment. Frequency stability of a high order is readily obtainable on frequencies up to about 200 Mc/s., and with the increased activity on VHF bands every effort should be made to obtain such stability to avoid interference with other transmissions. In addition, receiver selectivity is often as high as that used on the lower amateur bands and unstable transmissions will present reception difficulties. The use of simple oscillators is, therefore, not advised on frequencies below 200 Mc/s. On such frequencies the technique is very similar to that on lower frequencies, namely, a stable oscillator followed by a number of frequency multiplying stages. The stability of the oscillator becomes increasingly important as the output frequency of the transmitter is increased, since any variations in the oscillator frequency will be correspondingly multiplied. For example, using an oscillator on 7·5 Mc/s. in a 60 Mc/s. transmitter, a variation of 1 kc/s. in the oscillator becomes 8 kc/s. variation at the output frequency. Thus, crystal heating, voltage changes, etc., in a variable frequency oscillator as well as "chirpy" keying methods must be carefully avoided.

If satisfactory results are to be obtained careful choice of components and layout is essential. Parasitic oscillations are likely to be of the same frequency order as the desired output, and suppression of these parasitics may also suppress the wanted frequency, hence the desirability of arranging the circuit so that parasitics do not occur.

By-passing

Among the points to be watched is the selection of non-inductive condensers for by-pass use. It is usually better to choose small mica, silvered mica or ceramic condensers of low capacity rather than bulkier, higher capacity condensers. If a large capacity is essential, then it may be advantageous to connect a small non-inductive condenser in parallel with the larger one. All leads, whether in tuned circuits or by-pass circuits, should be as short as possible, and there should be one common earthing point for each stage. Where cathode bias is used, by-passing should be to cathode rather than chassis, but unless a really high quality by-pass condenser is available for the cathode circuit, the use of cathode bias is not advised as it is liable to introduce a common impedance into both grid and anode circuits.

Valves

Further, if best results are to be obtained only valves designed for use at very high frequencies should be employed. Some suitable valves with their main characteristics are given in Table III. Base connections are given in Table IV. Further details of British valves in this list are obtainable from the

TABLE III.

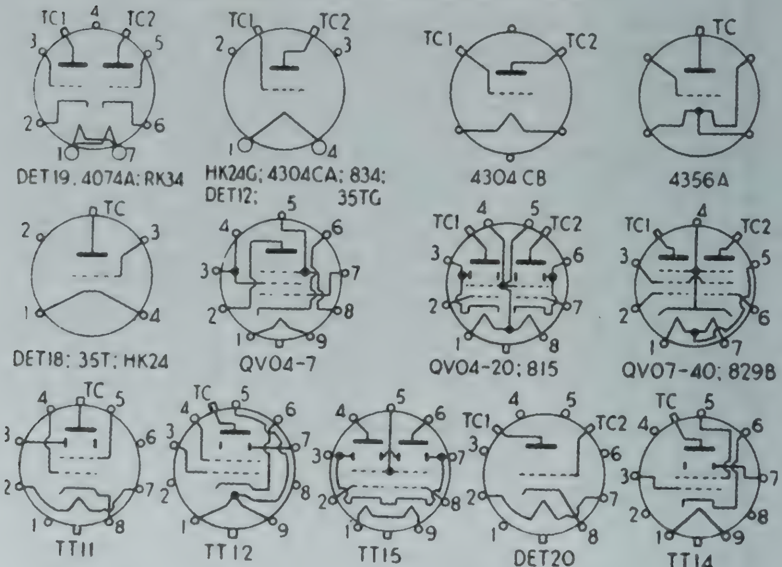
Details of some VHF Transmitting Valves.

Manufacturer.	Type No.	Class.	Cathode. V.	Cathode. A.	Max. Anode Volts.	Max. Anode Current mA.	Max. Anode Diss. Watts.	Max. Screen Volts.	Max. Grid mA.	Drive Power Watts.	Output Power Watts.	Max. Freq. Mc/s. (full ratings).	Max. Freq. (reduced ratings).
S.T. & C. .. M-O .. Raytheon ..	4074A .. DET19 .. RK34 ..	Twin Triode	6.3	0.8	300	2 × 50	2 × 5	—	18	1.8	16	100	300
M-O ..	TT14 .. TT12 ..	Tetrode ..	6.3 19	1.27 .42	600	100	25	300	5	0.4	40	60	150
S.T. & C. .. Mullard .. R.C.A. .. M-O ..	4304CA-CB TY1-50 .. 834 .. .. DET12 ..	Triode ..	7.5	3.3	1,250	100	50	—	27	—	90	100	300
S.T. & C. ..	4356.. ..	Triode ..	5	5	1,500	120	50	—	—	—	88	100	250
Eimac .. M-O ..	35T .. .. DET18 ..	Triode ..	5	4	2,000	150	70	—	35	30	225	60	150
H. & K. ..	HK24 ..	Triode ..	6	3	2,000	75	25	—	25	4.5	100	60	—
Mullard ..	QVO4-7 ..	Beam Tetrode	6.3	.61	300	50*	7.5	250	—	—	6.3	150	—
Mullard .. R.C.A. ..	QVO4-20 .. 815	Twin Beam Tetrode ..	6.3	1.6	400	—	2 × 10	225	5	0.2	30	240	—
M-O ..	TT15 ..	Twin Tetrode	6.3	1.6	300	2 × 45	2 × 7.5	200	2 × 3	0.2	20	60	150
Mullard .. R.C.A. ..	QVO7-40 .. 829B ..	Twin Beam Tetrode ..	6.3	2.25	750	2 × 120	2 × 20	225	2 × 7.5	0.8	87	250	—
M-O .. R.C.A. .. S.T. & C. ..	KT8C .. 807 .. .. 5B/250A ..	Beam Tetrode	6.3	1.27	600	95	25	300	6	0.5	38	60	125
M-O ..	DET20 ..	Triode ..	6.3	0.2	300	25	3.5	—	—	—	4	480	—
M-O ..	TT11 ..	Tetrode ..	6.3	0.8	300	50	7.5	250	3	0.1	6.7	100	150
R.C.A. ..	832	Twin Tetrode	6.3 12.6	1.6 .8	500	2 × 40	2×7.5	250	2.5	0.2	25	100	150

manufacturers. For most economical results the use of beam tetrodes is recommended as the drive required is very small and hence low power driving stages will suffice. As a typical example, the *Mullard* QV04–20 requires only 0.2 watts grid drive to give 30 watts RF output as a Class C amplifier For those who prefer triodes a number of suitable valves are available but somewhat more drive is required.

TABLE IV.

Base connections of Valves listed in Table III.



Neutralising

With careful layout and efficient screening of grid and anode circuits, many of the beam tetrodes will operate as power amplifiers on 60 Mc/s. without neutralising, but as frequency is increased or if screening is not adequate, then neutralisation becomes necessary. The anode-grid interelectrode capacity is very small in these valves, being less than ·2 $\mu\mu$F in some cases, and hence normal neutralising condensers are useless. A simple method, particularly suitable to twin beam tetrodes, is to use the anode of the valve as one side of the neutralising condenser, and a length of 12 or 14 SWG copper wire, connected to the appropriate end of the grid circuit, as the other side of the condenser. By altering the spacing between the wire and the anode, the capacity can be adjusted. With the grid circuit under the chassis and anode above *Denco* feed-through insulators will be found useful. With an RCA815 using 4″ of wire above the chassis the correct spacing between the neutralising wire and the valve envelope will be found to be about 1″. Some modification of this method is required for single-ended stages, the neutralising wire requiring to be placed close to the end of the tank coil remote from the

anode. A small metal disc attached to the wire can be used to increase the capacity if desired. These methods are shown diagramatically in Fig. 21.

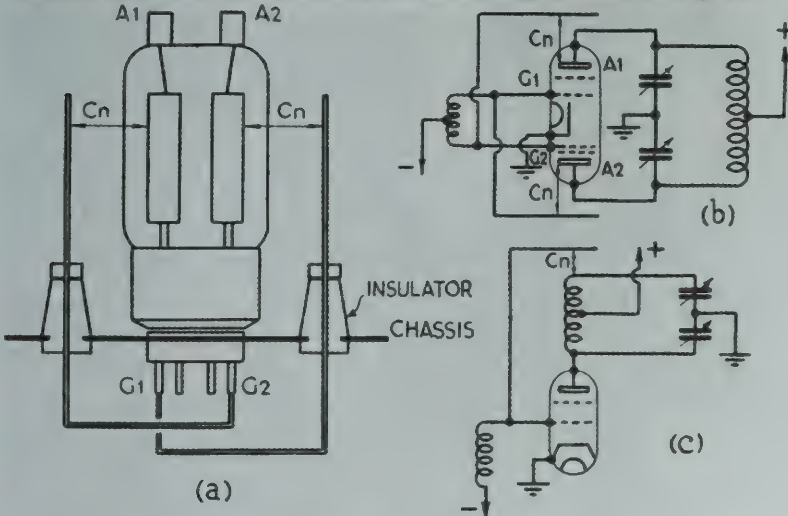
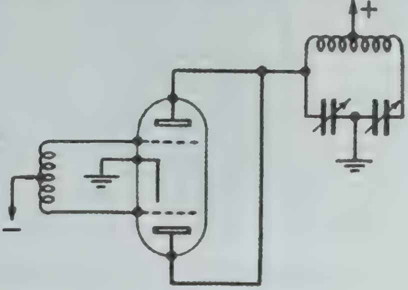


Fig. 21.
Method of neutralising beam tetrodes. (*a*) and (*b*) twin tetrodes, (*c*) single tetrodes.

Frequency Multipliers

Most of the valves shown in Table III can be used as frequency multipliers, the bias and drive being increased for efficient operation. The DET19 and its equivalents are useful as push-push doublers (see Fig. 22). Two QV04–7's or two 807's can be used in a similar circuit with good efficiency and the aerial can be fed direct from such a stage if desired, although the use of a straight power amplifier is advised whenever possible. The QV04–20 and QV07–40, (Fig. 23), with their equivalents make good treblers, while the QV04–7 works well as a doubler, trebler or quadrupler. For the lower frequency stages of the transmitter, valves normally used at those frequencies, *e.g.* 6F6, 6V6, 6L6, can, of course, still be employed, trebling and quadrupling replacing the more usual doubling in order to reduce the number of stages required.

Fig. 22.
Push-push doubler, suitable for use with DET19 or RK34, etc.



Tuned Circuits

At 60 Mc/s. ordinary coil and condenser tuned circuits (as shown in Fig. 24) are reasonably efficient and, due to their compactness, are usually employed in preference to resonant lines, but as frequency is increased the losses due to radiation and the difficulty of obtaining a reasonable sized coil for the normal type of tuned circuit give an increasing advantage to the linear circuit. In this type two conductors—usually $\frac{1}{4}''$ or $\frac{1}{8}''$ tubing or even large diameter copper wire—are run parallel to each other and spaced about their own diameter. The circuit is tuned by adjusting the lengths of the conductors by means of a shorting bar, the D.C. input being made at the same point. The effect on efficiency of using linear tank circuits is well shown by the curves of Fig. 25 which indicate the efficiency of the *Standard Telephones* 4356A triode used as a Class C oscillator on various frequencies. At 100 Mc/s. both coil and line circuits give 66% efficiency, but at 200 Mc/s. the coil circuit efficiency has dropped to 46% while the line circuit still has an efficiency of 62%. This also means that the highest frequency at which the valve can be used is greater with a line circuit than with coils.

The lengths of the lines required for resonance are $\frac{1}{4}\lambda$ but due to the length of leads inside the valves and to the interelectrode capacities the actual length in practice is less than this figure. Using twin beam tetrodes, such as

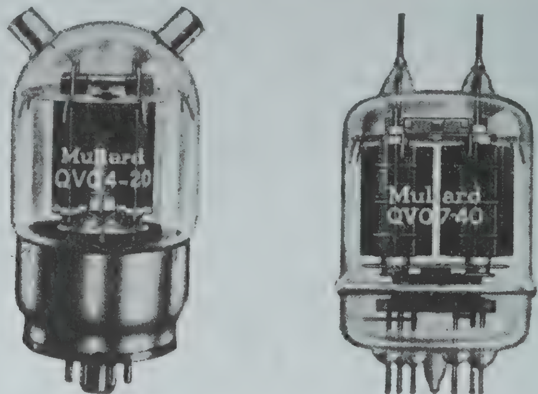


Fig. 23.
The Mullard family of VHF beam tetrodes, QV04-7, QV04-20 and QV07-40.

the QV04–20, the length required is about 15″ on 144 Mc/s. The lengths will, of course, be slightly affected by the size of conductor used and the spacing.

An alternative to the adjustable shorting bar method of tuning is to use a small, low-loss, tuning condenser at the open end (*i.e.* anode end). With this method rather shorter lines will be required due to the extra capacity. High-quality neutralising type condensers are suitable for this function. Resonance

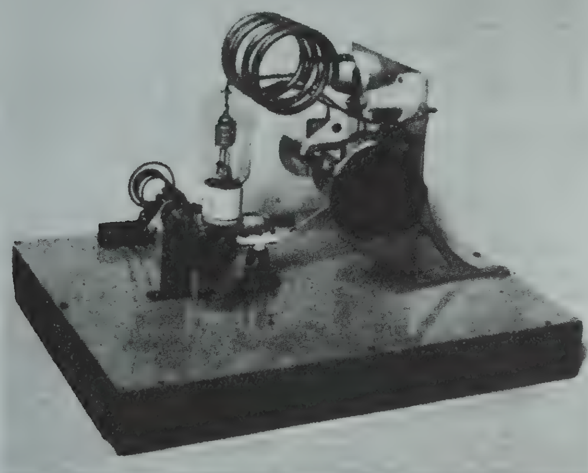


Fig. 24.
A 60 Mc/s. power amplifier using an Osram DET18 or Eimac 35T in use at G2WS.

can be determined by the usual means, such as dip in anode current or striking of a neon lamp, etc. Aerial coupling to such a circuit can be by tapping on to the lines *via* high-voltage type condensers, or better, by means of a coupling loop, hairpin shape, placed near to the shorted end of the lines. This is shown in Fig. 26.

In coil type circuits it is possible to employ a larger coil by using the anode-cathode capacity of the valve in series with the tuning condenser, instead of the more usual parallel arrangement, as is shown in Fig. 27.

Three Valve Crystal Controlled Transmitter for 60 Mc/s.

Fig. 28 shows the circuit diagram of a three valve transmitter designed for use on frequencies around 60 Mc/s. It can, however, be readily adapted for use on frequencies up to 144 Mc/s. by employing a higher frequency crystal and/or increasing the frequency multiplication of the second stage from two to three. By adding a further stage operation up to 200 Mc/s. is obtainable.

The two driver stages use beam tetrodes (*Mullard* type QV04–7) giving ample output to drive the power amplifier (*Mullard* QV04–20 or RCA815), to 25 watts or more.

The first stage is a quadrupling tritet, using a 7·5 Mc/s. crystal. With an

anode voltage of 160, the anode and screen currents of the QV04-7 total about 20 mA., and under these conditions crystal heating is negligible and the frequency is stable even after multiplication by eight. If the anode voltage is raised to 300 (the maximum permissible) this is no longer true and frequency drift becomes noticeable. Since adequate output on 30 Mc/s. is obtained with the lower voltages it is recommended that these be used. Parallel feed to the anode is recommended but there is no objection to the use of the more usual series feed.

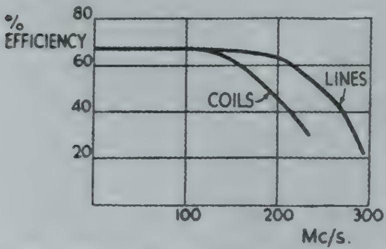


Fig. 25.
Variation of efficiency of S. T. & C. 4356A triode, with frequency when using lines and coils in the tuned circuit. *(S. T. & C. Ltd.)*

The second QV04-7 doubles the frequency to 60 Mc/s. As appreciable third harmonic output is obtainable with this valve, care must be taken to ensure that the correct harmonic is selected. Plate and screen current total 25 mA. when the anode DC voltage is 300 and the grid bias — 72 volts. For CW operation the screen supply to this stage is keyed, and it is for this reason that the resistance R5 is incorporated. Provided that grid bias is maintained at the figure given above, no "space-wave" is radiated, and the keying will be found to be free of "chirp." A simple filter of ·01 μF and 100 ohms across the key will ensure that no key clicks are radiated. No voltage stabilising device is employed; when the key is depressed the current drawn by V1 changes by less than 1 mA.

The grid circuit of V3, which is untuned, consists of a two-turn coupling coil pushed between the centre turns of the coil of the 60 Mc/s. circuit in the anode of V2. About 50 volts of negative bias are applied to the centre point of this coupling coil for the grids of V3. The drive can be adjusted by altering the coupling; with tight coupling it will be found that there is far too much drive. It is possible to drive the QV04-20 to 12 mA. grid current, which is three times the makers' recommended figure. Further adjustment is possible by varying the tuning of V1.

The screen grid is fed from a potentiometer connected between the modulated side of the modulation transformer and HT negative. This enables modulation of both plate and screen grid to be effected, and also minimises the rise in voltage on the screen grid when the key is up for CW

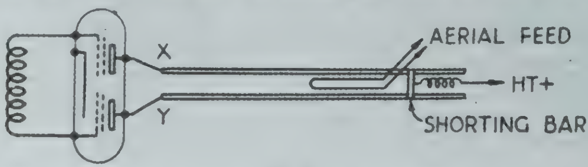


Fig. 26.
Linear tank circuit, suitable for use on 144 Mc/s. Tuning is by adjustment of position of shorting bar. Alternatively, a small variable condenser may be connected between X and Y.

operation. For 25 watts input the makers' figures of 325 volts on the anode and 165 on the screen grid will be found ample. For higher powers the anode voltage can be increased to 400.

No neutralisation was necessary in the original of this transmitter. All wiring and tuned circuits, with the exception of the PA output, were placed under the chassis so that screening was as complete as possible. Should it be found necessary to neutralise, the method described earlier in this chapter should be employed.

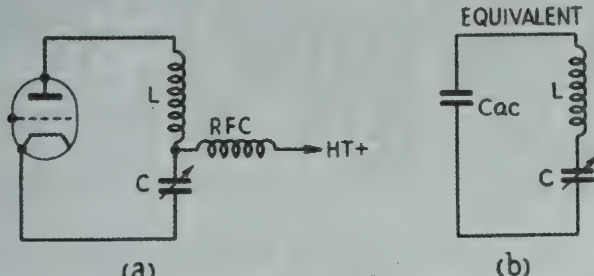


Fig. 27.
Showing how a large coil can be used at VHF by placing the anode-cathode capacity of the valve in series with the tuning capacity.

By substituting parallel lines for the output circuit, and increasing the bias on the QV04–20 to 120 volts, as well as increasing the drive, good output is obtainable on 180 Mc/s. By using an 8 Mc/s. crystal, and making V1 and V3 treble, while retaining V2 as a doubler, output can be obtained on 144 Mc/s. It is, however, advised that a further QV04–20 be added, when really good output should be obtained on this frequency. A linear tank circuit or parallel lines will be found to give superior results to the coil and condenser type.

Medium Power Transmitter for 60 to 70 Mc/s.

Fig. 29 shows a circuit which may be used on 60 to 70 Mc/s. with an input to the power amplifier of 80 to 100 watts. It may also be used as a driver unit for a high-power transmitter on 120 to 150 Mc/s. by converting the final stage to a doubler. There will be sufficient output to drive, say, two DET12's in push-pull, or alternatively, the output of the doubler may be fed direct to the aerial. Other output frequencies are available by changes in the frequency multiplication of one or more stages, and as with all VHF transmitters employing frequency multipliers, care must be taken to ensure that the output frequency from the final stage is that intended.

The oscillator stage, using a small audio frequency type pentode, is crystal controlled on about 7 Mc/s. Oscillation is assisted by the small feedback condenser C1. Satisfactory operation may be possible without this condenser, but results are more certain with it. The anode circuit is tuned to the crystal frequency and capacitively coupled to the second stage. Both second and third stages are treblers using KT8 or 807 types valves. Better output may be obtained by tapping the anode one or two turns up the coils. Small by-pass condensers should be connected across the heaters of the valves in the first three stages, one side of the heater wiring being earthed. This is not shown in the circuit diagram. A capacity of 500 $\mu\mu F$ is adequate.

The fourth stage can be operated as a straight power amplifier or as a

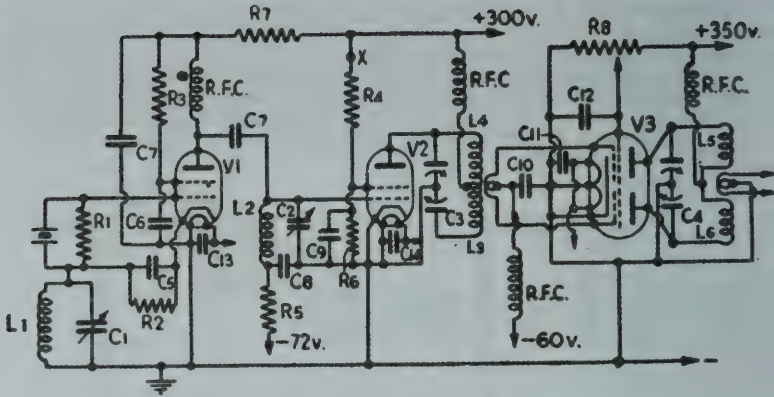


Fig. 28.
Three-valve crystal-controlled transmitter for 60 Mc/s., easily adaptable to higher frequencies.

R1	50,000 ohms 1 watt.	C1	60 $\mu\mu$F var.	C12	·01 μF in parallel with ·0005 μF.
R2	100 ohms ½ watt.	C2	20 $\mu\mu$F var.		
R3, 4	25,000 ohms 1 watt.	C3	15 + 15 $\mu\mu$F var.	C13, 14	·0005 μF.
R5	10,000 ohms 1 watt.	C4	20 + 20 $\mu\mu$F var.	V1	QV04-7.
R6	100,000 ohms 1 watt.	C5	·001 μF.	V2	QV04-7.
R7	6,000 ohms 5 watts.	C6, 7, 8, 9, 10, 11	·0005 μF.	V3	QV04-20.
R8	50,000 ohms var. 5 watts (see text).				

L1 9 turns, 24 S.W.G., close wound on 1" former.
L2 9 turns, 14 S.W.G., ¾" diameter spaced to occupy 1½".
L3 7 turns, 18 S.W.G., ½" diameter spaced to occupy 1¼".
L4 2 turns, interwound with L3 at centre.
L5 6 turns, ⅛" copper tube, 1" diameter spaced ⅛" between turns, arranged 3 and 3 with ½" gap at centre for L6.
L6 2 turns, 12 S.W.G., 1" diameter.

doubler. In the latter case the neutralising condenser should be adjusted to give greatest RF output. It may be found desirable to insert chokes in the filament leads to this stage at the points marked X. Such chokes must be capable of carrying the 3¼ amps. of filament current. It should be noted that this stage needs a separate filament supply from the other three stages. Cathode bias has been included in each stage, sufficient to limit the anode current to a safe value when the drive is removed, but battery bias can be used instead if desired.

In constructing this transmitter the general notes on layout at the beginning of this chapter should be borne in mind. When V4 is used as a power amplifier there must be adequate screening between the tuned circuits in the anodes of V3 and V4. Any parasitic oscillations occurring in stages V2 and V3 may be eliminated by inserting 50 ohm 1 watt resistors in the screen grid leads (*i.e.* between the valve terminal and the junction of the feed resistor and by-pass condenser.

Other Circuits

Many other circuit arrangements are, of course, possible and the two described here are intended mainly to show the type of valve line-up, etc.,

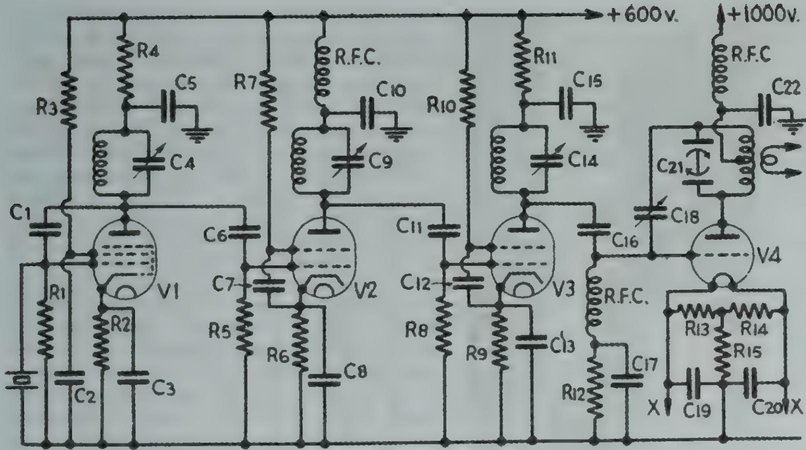


Fig. 29.
Transmitter circuit for 60 to 70 Mc/s., or driver unit for 120 to 150 Mc/s.

R1	50,000 ohms 1 watt.	R15	1,500 ohms 20 watts.	C17	500 $\mu\mu$F.
R2	250 ohms 1 watt.	C1	1 $\mu\mu$F Ceramic.	C18	4 $\mu\mu$F neutralising.
R3	80,000 ohms 1 watt.	C2, 7, 8,		C19, 20	·001 μF.
R4	10,000 ohms 10 watts.	10, 12, 13,		C21	25+25 $\mu\mu$F var.
R5, 8	100,000 ohms 2 watts.	15, 22	·005 μF.	R.F.C.	RF Chokes.
R6, 9	500 ohms 5 watts.	C3, 5	·01 μF.	V1	EL32.
R7, 10	40,000 ohms 2 watts.	C4	60 $\mu\mu$F var.	V2, 3	KT8C, 807.
R11	1,000 ohms 10 watts.	C6, 11, 16,	100 $\mu\mu$F.	V4	4304CB, DET12 or 834.
R12	20,000 ohms 10 watts.	C9	40 $\mu\mu$F var.		
R13, 14	100 ohms 1 watt	C14	20 $\mu\mu$F var.		

which will give results on frequencies up to 200 or 250 Mc/s. For operation on frequencies in the 144 to 146 Mc/s. band an 8 Mc/s. crystal can be employed followed by two trebling stages and one frequency doubling stage, the order being a matter to be decided by the valves available. Alternatively, a 6 Mc/s. fundamental frequency followed by quadrupling, trebling and doubling stages will produce the necessary output frequency.

Modulation

Modulation, to provide telephony transmissions, may be by any of the accepted methods. Due to the difficulty of providing adequate earthing connections for VHF the problem of keeping RF voltages out of the speech amplifier may be encountered. Two points on the chassis only a few inches apart may be at very different RF potentials. Large voltages may develop on the screening of screened microphone leads, etc., if these happen to be a half-wave long and earthing at more than one point may be necessary, as well as the inclusion of suitable chokes and by-pass condensers in the amplifier input. Grid stoppers are advisable in most, if not all, AF stages.

BCL Interference

For similar reasons, broadcast interference can be severe in the immediate

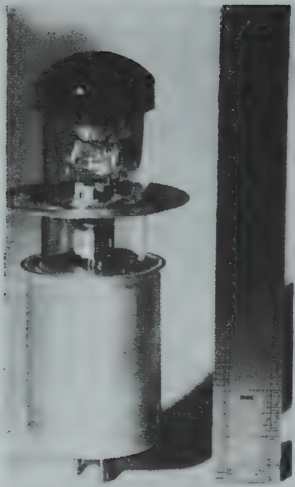


Fig. 30.
The 3B 401J double disc seal valve for operation up to 650 Mc/s.
(*S. T. & C., Ltd.*)

neighbourhood of a VHF transmitter. Disconnecting aerial and earth from the broadcast receiver may make no difference, and in all probability the setting of the volume control will have no effect on the interference. The trouble is due to the receiver chassis acting as a receiving aerial at VHF and the voltage variations produced between various points in it become applied to the audio stages and rectified. An RF choke in the grid of the first audio stage, as close to the grid terminal as possible will sometimes remedy this. Receivers using top-cap grid valves in the audio stages are particularly prone to trouble. The connection to the grid is usually made by means of a screened connector which may be 6 inches or more long. It is desirable that both ends of the screening should be connected to the chassis as directly as possible. In fact, all earthing and by-passing must be suitable for VHF operation. It may be found necessary to by-pass the mains leads to the chassis *via* 500 $\mu\mu$F condensers to make the cure complete, but the condensers on their own will not usually cure the trouble.

420 Mc/s.

Valves of normal design will not operate at frequencies much in excess of 300 Mc/s. This is due to electron transit-time effects and to the inductance of the leads inside the valves. At the time of writing the band of frequencies from 420 to 460 Mc/s. had not been made available to British amateurs, and with the necessity of a technique differing considerably from that used on lower frequencies, much experimental work will undoubtedly be required to evolve the most suitable circuits for use in amateur transmitters in this band.

Valves of the disc-seal type, sometimes known as "lighthouses," seem to present possibilities. Several types are made by *Standard Telephones* and *General Electric Co.*, from whom details are available. One of these, the 3B/401J, is shown in Fig. 30. Provided adequate ventilation is available, up to 40 watts anode dissipation is permissible. These valves are designed for use in concentric line circuits, but at 420 Mc/s. good results will probably

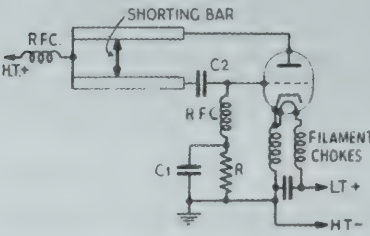


Fig. 31.
Suggested circuit for 420 Mc/s. Oscillator. The parallel lines can be of ¼" tube and about 5" long. Tuning is by adjustment of shorting bar position. R, 5,000 ohms, C1, 100 $\mu\mu$F, C2, 10 $\mu\mu$F.

be obtained with parallel line circuits. Fig. 31 shows a typical arrangement. The length of the parallel lines will be only a few inches, so the need for careful lay-out of the circuit will be evident. Stray capacities and any unnecessary length in connecting leads must be avoided at all costs. Because of the appreciable inductance of even short connecting wires, RF chokes, consisting of a few turns of suitable gauge wire, are connected in grid and filament leads to ensure that these electrodes are at the correct RF potentials.

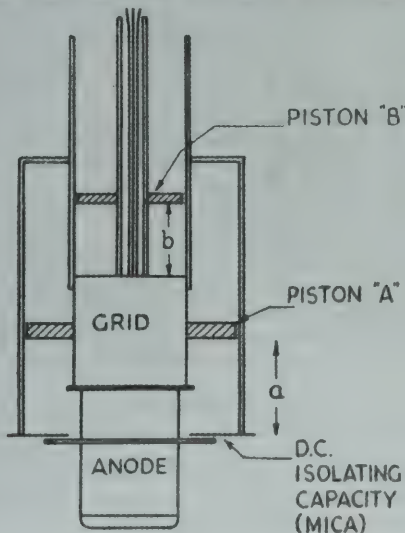


Fig. 32.
Disc seal triode in concentric line circuit suitable for use on 420 Mc/s.

For those desirous of using concentric tuned circuits, Fig. 32 shows such a circuit. A common tube forms the inner element of the anode-grid line and the outer element of the grid-cathode line. The anode is joined by a condenser to the outer tube of the resonator, so that no DC reaches the grid. Frequency is adjusted by a sliding piston (A) in the anode-grid line, and another (B) in the grid-cathode line to maintain resonance. Arrangements must be made, when constructing the oscillator, to adjust these pistons.

The ends of the filament and the centre tap are all connected through 100 $\mu\mu$F condensers to the centre conductor of the concentric system and leads are brought out from the filament and centre tap through the middle conductor in order to enable DC connections to be made.

RF power may be extracted by means of a pick-up loop inserted in one of a series of holes in the outer element of the anode-grid line.

CHAPTER 4 VHF RECEIVERS

INTRODUCTION — COMPONENTS — STRAIGHT RECEIVERS — SUPER-REGENERATIVE RECEIVERS — SUPERHETERODYNE RECEIVERS — THE DOUBLE SUPERHETERODYNE RECEIVER — SUPERHETERODYNE CONVERTERS

Introduction

VHF RECEIVERS may be grouped into five main categories:—

(1) Straight Receivers.
(2) Super-regenerative Receivers.
(3) Superheterodyne Receivers.
(4) Double Superheterodyne Receivers.
(5) Superheterodyne Converters.

The actual techniques of design and construction of these receivers depends on the frequency which is required to be received. Broadly speaking ordinary valve and circuit technique may be used at frequencies up to about 200 Mc/s., above this frequency special valves and circuits have to be used.

The performance of a VHF receiver depends to a large extent on the care and thought put into the mechanical layout, construction and choice of components. Only the best components should be used, rigidly mounted and laid out so as to keep lead-lengths as short as possible where RF currents flow.

Components

It is very important to choose only the best quality components for use in the RF sections of VHF equipment. Low Q dielectrics, such as ebonite and bakelite, should be avoided in coil formers, variable condensers and valveholders, where such materials may cause excessive losses and poor frequency stability with changes of temperature.

(a) Coils

Whenever possible, ceramic insulation should be used for coil formers where frequency stability is important, for example in superheterodyne local oscillator circuits, heterodyne wavemeters, transmitter oscillators, etc. For best results stable coils should be wound in grooves on ceramic formers. The windings should be made with bare or silver-plated copper wire, and wound under great tension. Under these conditions the finished coil will have a cyclic, or repeatable, temperature coefficient nearly equal to that of the ceramic former itself. In this way coils with a temperature coefficient of between 5 and 10 parts in 10^6 per degree Centigrade can be constructed. Coils wound on "distrene" or similar plastic material formers, and coils which are self supporting are not recommended where high frequency stability is required, although they are quite suitable in RF amplifier circuits and in super-regenerative receivers, which are of poor selectivity.

(b) Variable Condensers

Variable condensers for oscillator circuits of high stability should be carefully chosen to have low-loss insulation and a cyclic temperature coefficient. As it is almost impossible for the amateur to carry out measurements on variable condensers, the following points should be borne in mind when making a choice.

The condenser should be of sound, rigid mechanical design and of the same material throughout. Homogeneous construction eliminates relative expansion troubles. Single condensers are usually made all of brass with a single bearing; ganged condensers are usually constructed with a steel frame and a steel rotor shaft. Wide spacing of the condenser vanes is an advantage. The type of condenser illustrated in Fig. 33 is of all brass construction with ceramic insulation. Such condensers have a cyclic temperature coefficient of about +40 parts in 10^6 per degree Centigrade, and should not be confused with similar condensers, having distrene or similar plastic insulation, which have unpredictable thermal characteristics.

(c) Fixed Condensers

The choice of fixed condensers depends on the service for which they are required. Ceramic condensers, which usually have high temperature coefficients (−750 or +125 parts in 10^6 per degree Centigrade) should be

Fig. 33.
Stable variable condensers suitable for VHF applications. (Radiomart, G5NI (Birmingham) Ltd.).

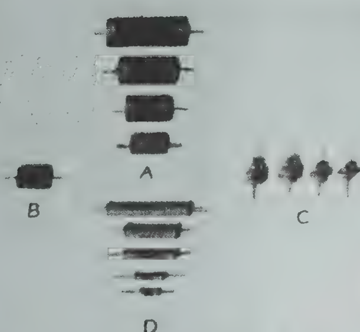


Fig. 34.
A group of different types of condensers as described in the text: A, Silvered mica condensers; B, Stacked foil type mica condenser; C, Cup and disc type ceramic condensers; D, Tubular ceramic condensers. (Dubilier Condenser Co. Ltd.).

Fig. 35.
Special VHF bypass condensers, chassis mounting and feed through types. (United Insulator Co. Ltd.).

avoided wherever they may influence oscillator stability, *e.g.* as grid condensers, padders, etc. A ceramic condenser of suitable size may, however, be used to compensate for the slightly positive temperature coefficient associated with coils and variable condensers to give an extremely stable oscillator combination.

Single-plate silvered mica condensers are recommended for use in the positions just mentioned. Such condensers have cyclic temperature coefficients of about 25 parts in 10^6 per degree Centigrade.

Ordinary stacked foil-type mica or ceramic condensers may be used as by-pass condensers in VHF equipment. The condenser lead-lengths should be kept short. It will be found that 500 $\mu\mu$F is a suitable capacity for use at 60 Mc/s. The best capacity for use at other frequencies may be chosen proportionally (*e.g.* 250 $\mu\mu$F for 120 Mc/s., etc.). Recently several special VHF by-pass condensers have appeared on the market. These are of ceramic or mica construction and provide a really low impedance by-pass to earth. Typical condensers of this type are shown in Fig. 35.

(d) HF Chokes

HF chokes for use in VHF equipment may easily be constructed by the amateur. A useful criterion for HF choke design, which appeared in the *Marconi Review*, July, 1945, is to take one-third of a wavelength of fine wire

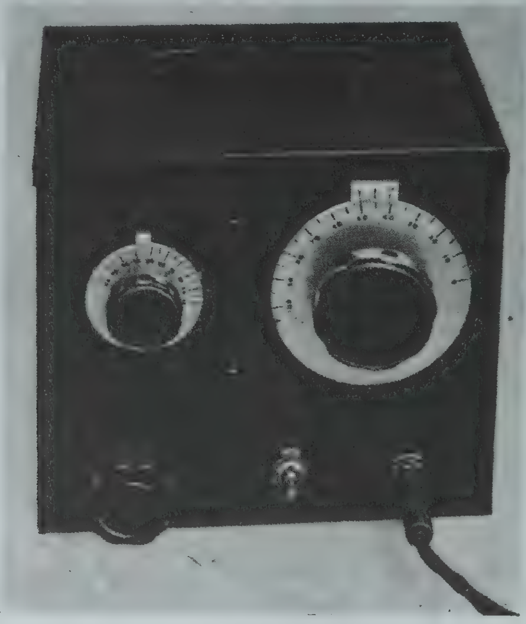


Fig. 36.
A 5-metre tuned radio frequency receiver.

and wind it to form a solenoid whose length is at least twice its diameter. It will be found quite convenient to wind such chokes on the outside of a 1 megohm ceramic tube-type ½ watt resistance.

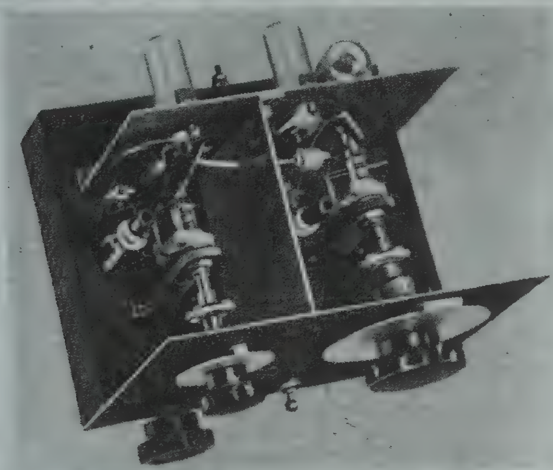


Fig. 37.
Photograph showing the layout of the principal components of the TRF receiver.

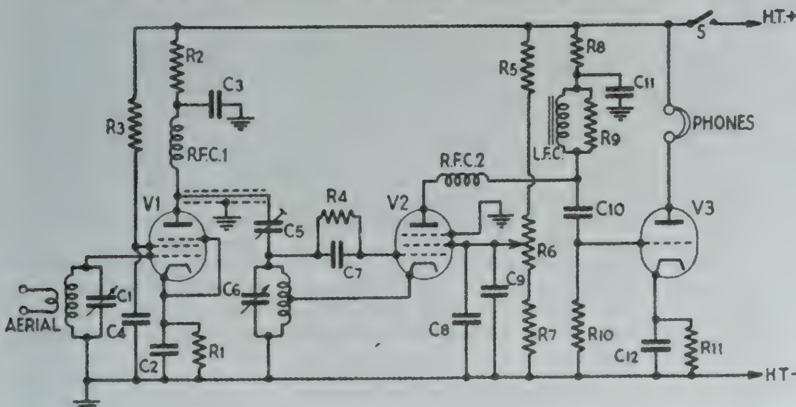


Fig. 38.
Circuit diagram of a tuned radio frequency receiver.

C1, 6	5 $\mu\mu$F.	C12	25 μF 25 v.	R7	10,000 ohms.
C2, 3, 4	·003 μF.	V1, 2	8D3 (Brimar)	R8	50,000 ohms
C5	3–15 $\mu\mu$F.	V3	6J5GT (Brimar).	R9	250,000 ohms.
C7	·0001 μF.	R1	150 ohms.	R11	3,000 ohms.
C8	·001 μF.	R3, 5	100,000 ohms.	RFC1, 2	RF Chokes.
C9, 11	4 μF.	R4	1 megohm.	Coils	See text.
C10	·01 μF.	R6	25,000 ohms.		

(e) Valve-holders

Valveholders should be of the ceramic or mica loaded distrene types especially when used in local oscillators.

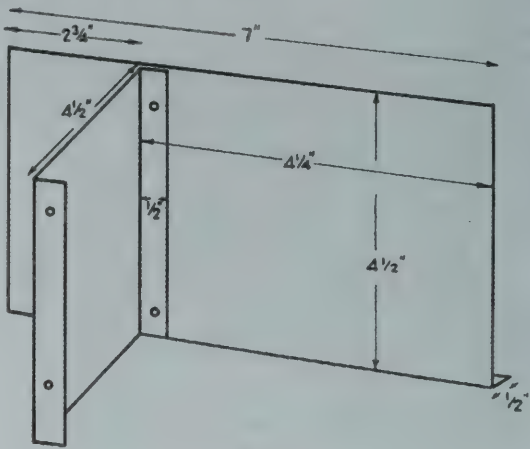


Fig. 39.
Constructional details of screens for TRF receiver.

Straight Receivers

A straight receiver is a receiver which employs some form of detector, usually with reaction, preceded by one or more stages of radio frequency amplification. Such receivers are well known in ordinary short-wave practice, and by careful design, construction and choice of components, may be made to operate quite satisfactorily up to 60 Mc/s. Straight receivers have the advantage of being simple and economical in the number of valves employed. The straight receiver, however, suffers from poor adjacent channel selectivity and is rather tricky to construct and adjust so as to have a smooth reaction control.

Straight receivers are seldom used in modern VHF practice because of their poor selectivity and difficulty of adjustment. They have a place in the amateur's shack because of their low cost. Straight receivers are not recommended, however, even for amateur purposes, at frequencies higher than 60 Mc/s.

A modern straight receiver covering 56–60 Mc/s. using the new seven-pin miniature valves is illustrated in Figs. 36 and 37. The circuit (Fig. 38) uses one RF amplifier stage followed by a regenerative detector and a triode LF amplifier. The circuit is conventional but care has been taken with the lay-out and construction to give maximum performance, smooth control and freedom from hand capacity effects (Fig. 39). The whole receiver, excluding power supply, is assembled in a metal cabinet, measuring 7" × 8" × 6½". The sub-panel and partition are made from 16 or 18 SWG aluminium sheet.

When wiring-up the signal frequency stages, care should be taken to return all earth connections to single points on the chassis, one for each stage.

The coils are wound on $\frac{3}{8}''$ diameter formers with 18 or 20 SWG wire, the ends of the coils being anchored to small screws set in the formers at suitable points. Both coils are wound with 7 turns of wire, the detector coil being tapped one turn from the earthy end. The aerial coil has a primary winding of $1\frac{1}{2}$ turns interwound at the earthy end of the 7 turn coil.

Smooth reaction should be obtained when the reaction control is rotated; any harshness on the theshold of oscillation may be reduced by altering the resistance across the LF choke, and the value of the detector grid leak.

The tuning of the RF stage is quite flat, in fact it will be found when listening between 58·5 and 60 Mc/s. that no adjustment is required.

Super-Regenerative Receivers

A super-regenerative receiver is a receiver using a super-regenerative detector with or without preceding radio frequency amplification. The principle of super-regeneration is outlined below.

The steady oscillation of the detector valve is interrupted by some means at a supersonic frequency. This stopping and starting of the oscillations is called "quenching" and the rate at which it is performed is called the "quench frequency."

The quenching of the detector oscillations may be performed either by means of a separate quenching oscillator or by choosing the valve's grid leak and condenser combination so that blocking of the oscillations occurs at a rate depending on the grid circuit time-constant.

Normally the oscillations are initiated by thermal noise-pulses present in the detector tuned circuit. The incidence of a signal on the detector, however, modifies the timing and/or duration of the bursts of oscillation, depending upon the amplitude of the received signal, thus giving rise to an audio signal in the output circuit in the case of reception of a modulated signal. A more detailed description of the modes of super-regeneration may be found in Terman's *Radio Engineers' Handbook*, pp. 662–664 (McGraw-Hill), which book also gives references to original papers on the subject.

In the logarithmic mode of operation, with separate or self-quench method of operation, the audio output waveform will be distorted, especially on deeply modulated signals. The logarithmic input–output characteristic has the effect of giving the super-regenerative receiver a measure of ignition noise suppression. The small change of audio output for large variations of input signal level is well known to those who have had experience with super-regenerative receivers. The curve shown in Fig. 40 shows the output of a typical logarithmic mode super-regenerative detector plotted against the relative input.

The super-regenerative receiver has always appealed to the radio amateur because of its extreme simplicity and ease of tuning. Super-regenerative receivers have been constructed to operate at frequencies up to about 600 Mc/s. using "acorn" type valves and "line" type tuned circuits.

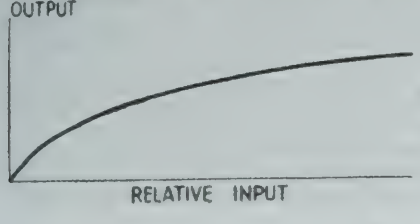


Fig. 40.
Super-regenerative receiver input/output characteristic.

The super-regenerative receiver, then, has the advantage of extreme simplicity and ecnomy in construction, high sensitivity, a measure of ignition interference suppression, flat tuning (enabling unstable signals to be held in tune) and a wide frequency coverage as evidenced by the well known "National 1-10" receiver, covering 30 to 300 Mc/s. with plug-in coils.

Such receivers, however, possess certain disadvantages which are becoming more and more serious as the VHF spectrum becomes more heavily populated. These disadvantages are: a lack of adjacent channel selectivity, a rather high noise level, inability to receive CW signals, and spurious radiation. The last mentioned disadvantage may be reduced, to within tolerable limits, by careful screening and the addition of RF amplifier stages, although it becomes difficult to design suitable RF amplifiers to operate at frequencies above 200 Mc/s. The frequency stability of the super-regenerative receiver is generally poor but as the set is at the same time very unselective this defect is not usually noticeable.

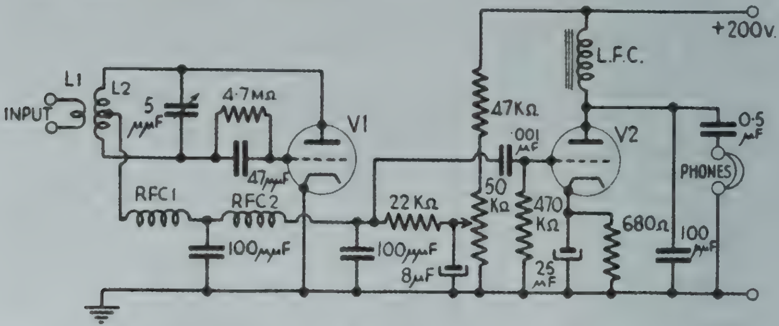


Fig. 41.
Circuit diagram of a simple VHF super-regenerative receiver.
V1. DET20, 9002, RL16, L63, etc. V2. L63. L2. 7 turns ½" dia. approx. for 60 Mc/s.

The circuit of a simple super-regenerative receiver which will work satisfactorily up to 300 Mc/s. is shown in Fig. 41. The detector stage uses an *Osram* type DET20 or some similar VHF triode connected in a self-quenching circuit. If the 60 Mc/s. band only is required the DET20 valve may be replaced by a valve of the L63 type. Super-regeneration is controlled by means of the 50,000 ohm potentiometer, and is adjusted by rotating this potentiometer until a loud rushing noise is heard in the headphones. Some experiment with the values of the choke (RFC1), the detector grid leak and the aerial coupling may be necessary before smooth super-regeneration is obtained.

The two chokes (RFC1 and RFC2) together with the by-pass condensers form filters to reject signal frequency and quench frequency currents from the LF circuits. A single L63 LF stage will give ample volume for a pair of headphones.

It is recommended that the receiver be constructed on an aluminium chassis and panel, care being taken to keep all RF and by-pass leads as short as possible. A simple set as described without an RF stage will give a good performance on signals as weak as 5 microvolts but the constructor is urged to be careful when using it to avoid interference, due to radiation, with amateurs and other VHF services which include air navigational aids, etc.

The disadvantage of radiation from a simple super-regenerative detector may be overcome to a large extent by the addition of one or two RF amplifier stages. A suitable circuit using the new seven-pin miniature valves is shown in Fig. 42. An RF pentode is used as an amplifier, followed by a triode super-regenerative detector. The screening of the detector should be as complete as possible, as indicated in the circuit diagram, if radiation is to be reduced to a minimum. Chassis by-pass condensers, similar to those shown in Fig. 35, may be used with advantage to filter all supply leads passing into the screened compartment.

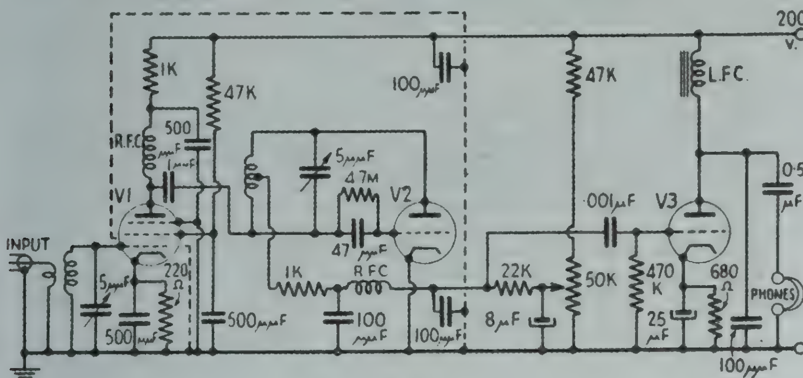


Fig. 42.
A super-regenerative receiver circuit with an RF stage to reduce radiation and improve selectivity.
V1. Z77, 8D3, CV138, ZA2, etc. V2. DET20, 9002, RL16, etc. V3. L63.

Superheterodyne Receivers

The superheterodyne is the most suitable type of receiver for general amateur use on the VHF bands. Its sensitivity, selectivity and good signal-to-noise ratio coupled with wide frequency coverage, extending to the shortest wavelengths yet explored, speak for themselves.

(a) Basic Principles

In a superheterodyne receiver an incoming signal voltage is mixed with a local oscillator voltage in a suitable mixing device, or frequency changer, producing an intermediate frequency signal which resembles the incoming signal in every way except that the carrier frequency has been changed. The fundamental advantages arising from this frequency change are that by careful choice of the intermediate frequency a high gain and a narrow bandwidth (good selectivity) may be obtained in the IF amplifier. A high gain is required in order to amplify signals which may be very weak, and a narrow bandwidth is required to give a high signal-to-noise ratio and good adjacen channel selectivity.

The basic superheterodyne circuit, in block diagram form, is shown in Fig. 43, where it can be seen that the signal and local oscillator voltages are fed into a mixing device. The exact form of the mixer depends to a great extent on the signal frequency. Modern triode-hexodes may be used up to

100 Mc/s. and pentodes up to about 200 Mc/s. Above this frequency, triode, diode and crystal mixers are used. Sharp cut-off pentodes are less noisy as mixers than triode-hexodes; triodes are quieter still.

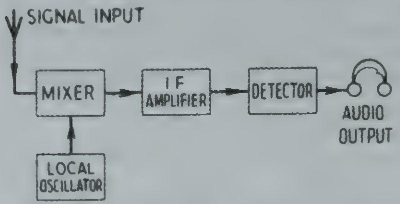


Fig. 43.
Basic superheterodyne receiver block diagram.

The mixer stage may or may not be preceded by an RF amplifier stage, depending upon the wavelength and availability of suitable amplifying valves. Pentode valves perform satisfactorily up to about 200 Mc/s. above which frequency neutralised grounded-cathode, or grounded-grid, triode valves may be used. The useful upper frequency limit with modern grounded-grid valves of conventional construction is in the region of 300 Mc/s. Special types of grounded-grid triode are, however, available for operation up to frequencies of several thousand megacycles per second. Either one or two RF stages are recommended where possible on account of the increase in signal-to-noise ratio that may be achieved.

Output from the mixer valve—being the difference frequency between the signal frequency and local oscillator frequency—is amplified by the intermediate frequency amplifier and then passed on to some form of demodulator or detector. The basic problem in superheterodyne design is to choose the various circuits which go to make up the units of Fig. 43, to fulfil the constructor's particular requirements.

(b) Design Requirements

The following list of requirements, which should be met in the design of a VHF superheterodyne for amateur purposes, is given as a guide.

(*i*) *Receiver Gain*

Receiver gain should be sufficient to receive signals at levels down to the inherent noise level of the RF stage. In the case of frequency modulation receivers the gain should be sufficient to saturate the limiter stage with the noise originating from the RF stage.

(*ii*) *Receiver Bandwidth*

The narrowest possible bandwidth should be used, consistent with sufficient width to cope with transmitter and receiver local oscillator frequency-drift. A narrow bandwidth makes for an increase in signal-to-noise ratio.

(*iii*) *Image Ratio*

There should be sufficient RF tuned-circuits to give a high image ratio using an intermediate frequency chosen with consideration of the requirements of stage gain and bandwidth. Three RF tuned circuits will give ample image rejection and two (*i.e.* one RF stage) will usually be sufficient. (See " Double Superheterodyne.")

(*iv*) *Tuning Control*

Bandspread tuning is preferable in amateur type receivers, covering just a little over each end of the band chosen. This arrangement simplifies ganging problems and makes possible a higher L to C ratio, giving greater stage gain.

(*v*) *AVC*

Automatic volume control is desirable and this should be capable of handling signals ranging from a few microvolts to one hundred millivolts without distortion.

(*vi*) *Noise Silencer*

In order to reduce the effects of ignition and other pulse types of interference a noise limiting circuit is usually considered essential in a receiver used in built-up areas or near main roads.

(*vii*) *LF Gain*

The low frequency gain of the receiver should be sufficient to operate a loudspeaker, or 'phones, at full strength on the signal level corresponding to the knee of the AVC curve.

(*viii*) *CW Reception*

The reception of CW Morse signals is possible at frequencies as high as 150 Mc/s. For this purpose a beat frequency oscillator may be incorporated.

(*ix*) *LF Response*

In the interests of obtaining the highest signal-to-noise ratio the audio bandwidth of the receiver should be restricted to approximately 300 to 3,000 cycles per second.

(*x*) *Tuning Scale*

An adequate directly calibrated tuning scale with a good slow-motion drive, free from back-lash, is of prime importance.

(*xi*) *Aerial Trimmer*

An aerial trimmer, mounted on the control panel and connected in parallel with the first RF stage tuned circuit, is an advantage in compensating for different aerials and the variation of aerial reactance over the band.

(*xii*) *Local Oscillator*

The local oscillator should be designed to have a high degree of frequency stability.

The extent to which these requirements can be met depends largely on economic resources and the individual amateur's constructional facilities, scientific knowledge and skill.

(c) Choice of Intermediate Frequency

The choice of intermediate frequency for use in a VHF receiver is a compromise between the requirements of a narrow bandwidth for selectivity and good signal-to-noise ratio, a high IF for good image ratio, and a wide bandwidth to cope with frequency drift. It has been found in practice that an IF of 1·6 Mc/s., using coils with a Q of about 100, critically coupled in pairs, gives satisfactory results at 30 Mc/s. The figure of 100 is found to be the Q of solenoid coils of economical size and construction over a wide range of frequencies. Since the Q of the circuits and the mid-band frequency fixes the bandwidth of critically coupled IF transformers it will be seen that the bandwidth of practical transformers is roughly proportional to the mid-band frequency. Frequency drift is found to be approximately proportional to signal frequency thus giving the useful criterion that the intermediate frequency should be chosen proportional to the signal frequency, using a 1·6 Mc/s. IF for a 30 Mc/s. signal frequency as a basis.

When a VHF receiver is required to cover more than one of several widely spaced frequency bands, some further compromise must be made, *e.g.* an IF of about 5 Mc/s. could be used for signal frequencies between 30 and 150 Mc/s. An IF of 10 Mc/s. would be more suitable for a receiver operating in the 150 to 300 Mc/s. band.

(d) The IF Amplifier

The IF amplifier of a VHF receiver is called upon to provide the major portion of the overall gain of the set as well as the required degree of adjacent channel selectivity.

It is common practice to employ two or three IF amplifier stages in a VHF superheterodyne receiver using pairs of slightly less than critically coupled circuits in the IF transformers. Circuits which are over-coupled should be avoided as they are difficult to tune-up. It may be shown that the gain of an IF stage depends only on the bandwidth and tuning capacity. Now as the bandwidth is fixed by such matters as oscillator stability, ease of tuning, etc., the gain per stage may only be increased by reducing the tuning capacity. Here again a compromise must be sought if the effects of detuning, due to change of valve input capacity with AVC bias voltage, are to be avoided. A further limitation of stage gain is caused by instability due to feed-back through the IF valve grid-anode capacity. A total tuning capacity of not less than 75 $\mu\mu$F (including strays) is recommended for use in amateur practice in IF amplifiers up to 12 Mc/s.

Fixing the bandwidth and tuning capacity in this way fixes the dynamic impedance of the IF transformer and the stage gain which may be expected. The theoretical stage gain is seldom achieved in practice in multi-stage amplifiers owing to feed-back. A stage gain of 20 to 30 times may be considered very satisfactory in a three stage amplifier operating at 5 Mc/s. using KTW61 or EF39 type valves. Higher gains can be obtained using EF50 type valves with very careful screening, but the absence of the variable-*mu* characteristic may introduce distortion.

Great care should be taken with the lay-out, screening, and by-passing of the IF amplifier stages in a VHF receiver. Grid and anode leads must be kept short and by-pass condensers chosen to provide a low impedance path to earth near the point where the cathode of the valve is earthed. Mica condensers with a capacity of ·01 μF will be found satisfactory for use as by-passes in IF circuits operating between 2 and 10 Mc/s. If possible the IF amplifier should be laid out in a straight line, and on no account should the output of the IF " chain " finish near the mixer stage, which is at the input end of the amplifier.

(e) A Practical IF Transformer Design

There are several IF transformers on the market suitable for use in amateur VHF superheterodyne receivers, but they can easily be constructed at home by the amateur himself. The following design is for a 5 Mc/s. transformer, suitable for use in a receiver covering 30 to 150 Mc/s.

The primary and secondary windings consist of 30 turns of No. 36 SWG enamelled wire, close-wound on a ½" diameter polystrene or distrene former. Both coils are wound in the same direction, the spacing between the windings being ⅝". It is important that the two inner ends of the coils should be made the earthy connections (*i.e.* HT and GB -) otherwise the coupling factor will be altered. Fig. 44 illustrates an amateur constructed transformer made to the above specifications. The circuit and details are given in Fig. 45.

Mullard 3-30 $\mu\mu$F concentric trimmers in parallel with 68 $\mu\mu$F silvered mica condensers make up the tuning capacity. Ceramic fixed condensers and compression-type trimmers should be avoided as both are prone to large capacity changes with temperature, giving rise to detuning troubles.

(f) RF and Mixer Stages

The design of the radio frequency section of a VHF receiver for amateur purposes depends very largely upon the valves which are available. There are several valves and many components which can be made to operate

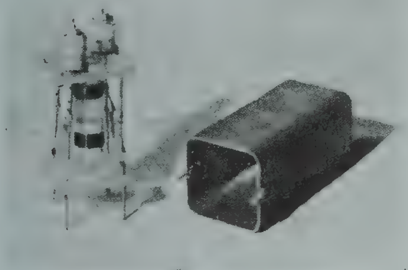


Fig. 44.
Photograph of home constructed 5 Mc/s IF transformer.

satisfactorily up to about 200 Mc/s., and this frequency may be regarded as the practical upper limit at which amplification can be obtained using easily procurable valves and straightforward circuit technique.

At frequencies above 200 Mc/s. it is common practice to use sets with either grounded-grid triode RF amplifier stages or no RF amplifier stage at all. In the latter case the signal is fed straight into a suitable diode or crystal mixer circuit. There is no definite dividing line between the two techniques although there is an overlap region between 150 and 250 Mc/s.

(i) *RF and Mixer Stages Below* 200 *Mc/s.*

The *Mullard* EF50 and *Osram* Z90 valves stand out "as-maids-of-all-work"

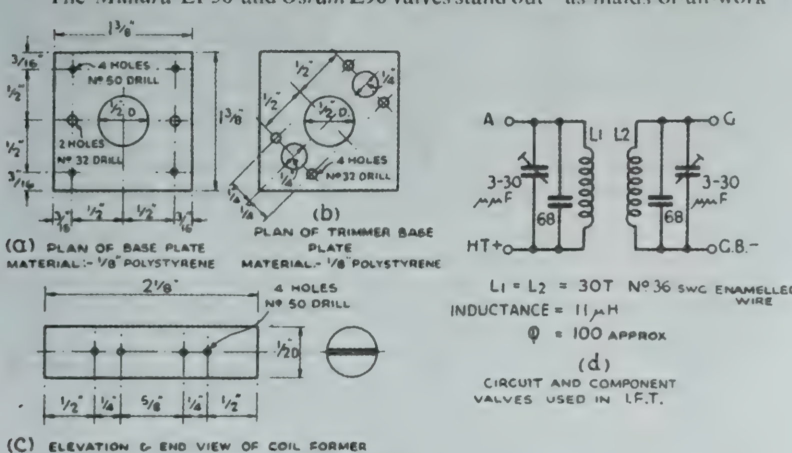


Fig. 45
Constructional details of IF transformer.

for VHF equipment up to about 200 Mc/s. Valves of this type have been produced in enormous quantities for Service applications and are readily obtainable. The Service equivalent type numbers are: R.A.F., VR91; Army, ARP35, CV1091 or CV1578. The *Mullard* RL7 or EF54 (Service type VR136 or CV1136) possesses an advantage over the EF50 in that it has a lower noise level. Additionally, by using four cathode leads improved performance at the higher frequencies is obtained.

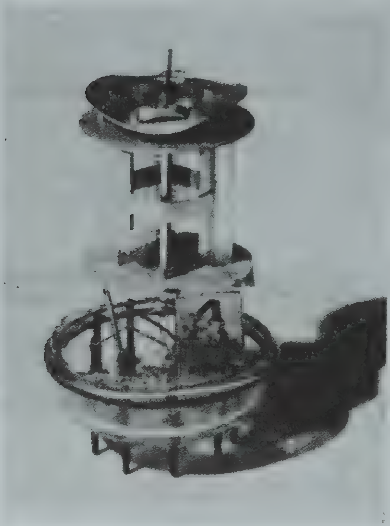


Fig. 46.
The ring seal type of VHF pentode.
(M.O. Valve Co. Ltd.).

Both these valves are steep slope RF pentodes mounted on a pressed glass (B9G) base, and covered externally by an aluminium screening can. The ring-seal construction, well illustrated in Fig. 46, ensures that the leads from the valve pins to the electrode structure are short and, therefore, have low inductance. EF50 and EF54 valves can be used as RF amplifiers without any trouble providing that care is taken to shield the input and output circuits. The pin connections are conveniently arranged so that a simple under-chassis screen is all that is required. The disposition of the screen in each case can be seen in Fig. 47. A practical tip, of interest to home constructors, is that tinplate is a very handy material for the construction of VHF chassis and screens. It is easily cut-up and the screens may be soldered in position with an ordinary electric soldering iron.

Although the two valves already mentioned stand out as being the commonest valves suitable for VHF work, there are several others available. For example the B9G series of valves is being superseded by a new range of miniature valves mounted on a button type (B7G) glass base. The *Osram* Z77

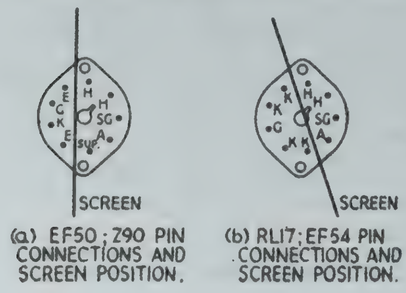


Fig. 47.
Diagrams showing the correct positions of the under-chassis screens for the EF50 and EF54 type valves.

is an example of a miniature valve which has an improved performance over the older EF50/Z90 types. Acorn type valves have fallen into disfavour owing to their fragile construction, although they still retain some electrical advantages at the upper limits of the VHF spectrum. Table IV lists most of the types of valve which are available for service between 30 and 300 Mc/s.

TABLE IV.

Valves Suitable for use in the RF and Mixer Stages of VHF Receivers.

Valve.	Type.	Base.	Upper Frequency (approx.) Mc/s.	Maker.
EF50	Pentode	B9G	200	Mullard
Z90	Pentode	B9G	200	Osram
VR91 .. ARP35 CV1901 CV1578	Pentode (as EF50)	B9G	200	Service Types
RL7/EF54 ..	Pentode	B9G	250	Mullard
VR136.. .. CV1136 ..	Pentode (as RL7)	B9G	250	Service Types
Z77	Pentode	B7G	250	Osram
8D3	Pentode	B7G	250	Brimar
CV138.. ..	Pentode	B7G	250	Service Type
9003	Var-mu Pen	B7G	250	American
9001	Pentode	B7G	250	American
6AK5	Pentode	B7G	250	American
954	Pentode	Acorn	300	American
ZA2	Pentode	Acorn	300	Osram
CV139.. ..	G.G. Triode	B7G	250	Service Type
EC91	G.G. Triode	B7G	250	Mullard
CV66	G.G. Triode	B9G	250	Service Type
6J4	G.G. Triode	B7G	250	American
3A/146J ..	G.G. Triode	Special	450	S. T. & C.

Fig. 48 illustrates a well tried circuit employing an RF stage and mixer for a VHF superheterodyne. The aerial is link-coupled to the first tuned circuit which in turn is coupled tightly to the grid of the RF valve. The aerial circuit will be heavily damped by the aerial and the valve input resistance and so will not provide a great measure of selectivity; tight coupling is desirable, however, for the best signal-to-noise ratio. For the optimum signal-to-noise ratio the aerial should be coupled a little tighter to the tuned circuit than is required for correct impedance matching. As under these conditions little selectivity is provided it may be desirable to use two RF stages.

The following constructional hints may help in giving trouble-free operation. All leads should be kept short and earth returns made direct to the chassis near the point where the cathode of the valve is earthed. Small mica condensers, such as the *Dubilier* type 635, 500 $\mu\mu$F, are suitable for by-pass purposes at frequencies from 30 to 100 Mc/s. Smaller capacities may be found desirable at the higher frequencies. Special VHF by-pass condensers

designed to have a very low inherent inductance are now available, and samples of these are illustrated in Fig. 35. The method of fixing will be obvious from the photograph.

The RF choke shown in the anode of the RF stage of Fig. 48 should be wound to the criterion given earlier in this chapter (pp. 38–39). For 60 Mc/s. work 65 inches of No. 38 SWG enamelled wire, wound on a ceramic type half-watt *Erie* resistance to take up the whole length of the resistance, will be found satisfactory.

The RF stage is capacity-coupled to the mixer valve grid circuit. An EF50 is shown again as a mixer valve, the signal and oscillator voltages being fed into the control grid. Experience has shown that the optimum condition for this valve as a mixer is with a grid bias nearly at cut-off (−4 volts approx.) and sufficient local oscillator injection to the control grid so that the valve

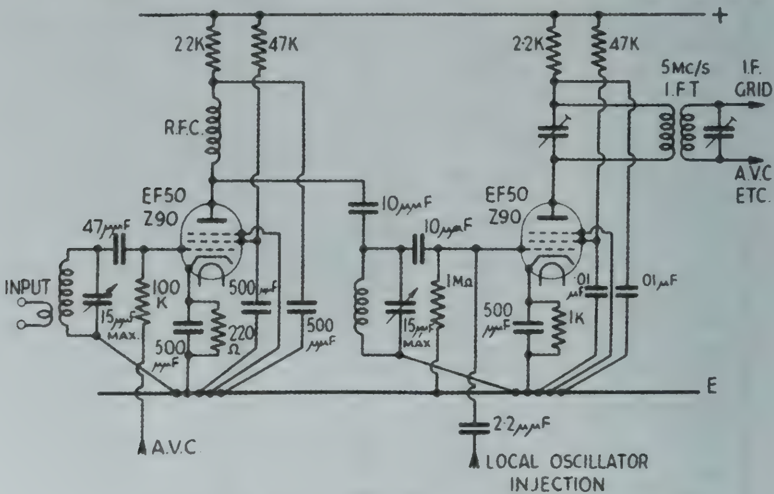


Fig. 48.
A well tried RF and mixer circuit for operation between 30 and 150 Mc/s.

is just short of taking grid current. The onset of grid current may be observed by inserting a microammeter in series with the earthy end of the grid leak and noting when the meter first shows a reading as the injection is increased. The injection should then be reduced to a figure slightly below this value.

It should be remembered that local oscillator and signal frequency components are present in the anode current of a mixer valve in addition to the IF component. With this in mind the anode and screen by-pass condensers on the mixer valve should be chosen to have a low impedance path to earth for VHF currents as well as IF currents. In practice this means that the capacity of these condensers should be a compromise between the IF and VHF requirements and should be returned to the chassis near the cathode earth point by the shortest possible leads.

The method of oscillator injection shown in Fig. 48 has been found satisfactory when a 5 Mc/s. IF or higher, has been used; no excessive "pulling" being evident. Other possible methods of oscillator injection are shown in

¶ Fig. 49. Suppressor grid injection has the disadvantage of requiring a high oscillator voltage, whilst cathode injection sometimes causes IF instability.

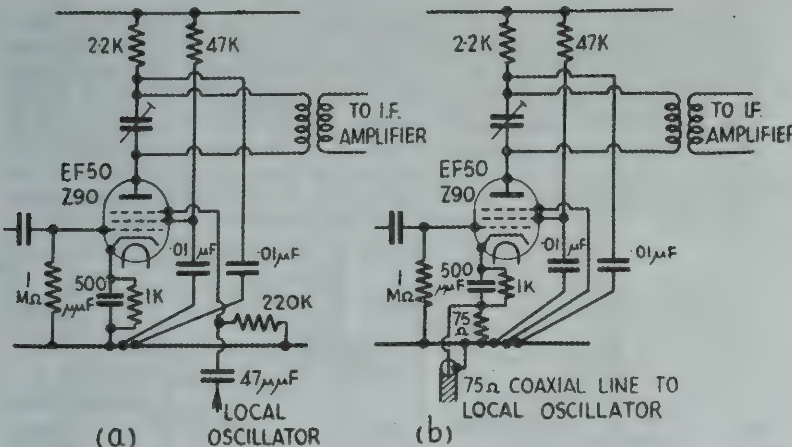


Fig. 49.
Alternative VHF mixer circuits, (a) Suppressor grid injection, (b) Cathode injection.

A recent advance in VHF receiver technique has been the introduction of the grounded grid, or common grid, triode amplifier RF stage. Such an amplifier may be made to operate as high as 3,000 Mc/s., using special valves and up to about 300 Mc/s. using valves of conventional construction, such as the CV66 and the CV139 (EC91). The great advantage of the grounded-grid triode is that it has a much lower equivalent noise resistance than a pentode valve, *e.g.* the CV66 is only half as noisy as the RL7 (EF54) which is one of the best pentodes from the noise standpoint.

The input impedance of a grounded-grid stage is very low, of the order of a few hundred ohms, and a wide band input circuit as shown in Fig. 50 may be used to couple it to an 80 ohm aerial feeder. This may be an advantage as no ganging troubles are introduced although, of course, the input circuit

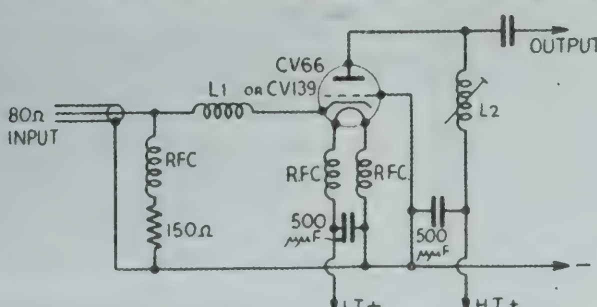


Fig. 50.
The grounded grid amplifier circuit.

provides little second channel rejection. It is good practice, therefore, to follow the grounded-grid stage by a stage of pentode amplification where possible, using selective circuits before feeding into a mixer.

The circuit of a CV66 grounded-grid amplifier stage suitable for 60 Mc/s. operation with mixer and local oscillator is shown in Fig. 51. The inductance in the anode circuit of the amplifier stage is pretuned to the middle of the band by a copper slug—the local oscillator being the only variable tuning control brought out on the front panel. The CV139 grounded-grid triode may be used in a similar circuit, and layout sketches for both CV66 and CV139 grounded-grid stages are given in Fig. 52. The circuit shown in Fig. 51 may be used as a 60 Mc/s. converter, output from the link of L4 being fed into a short wave receiver tuned to an IF of 5 Mc/s. By suitably proportioning the coils, the converter may be made to operate up to 150 Mc/s.

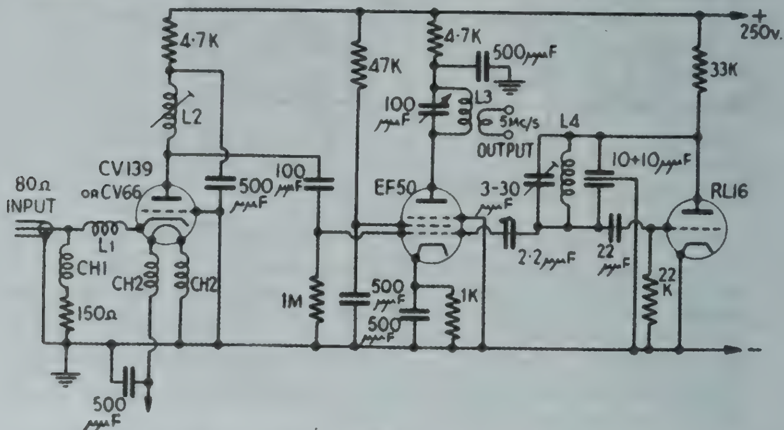


Fig. 51.

Grounded grid amplifier, mixer and local oscillator circuit for 58 to 60 Mc/s. operation.
L1. 10 turns No. 20 SWG ½" dia. L3. 5 Mc/s. IF coil.
L2. 6 turns No. 20 SWG ½" dia. L4. Oscillator coil to cover 53 to 55 Mc/s.

(ii) RF and Mixer Stages above 200 *Mc/s.*

The transition from ordinary coil and condenser circuits to line circuits takes place in the region between 150 and 250 Mc/s. The trend at the higher frequencies is to use crystal diode mixer circuits with or without grounded-grid RF stages. Capacity loaded line type circuits are commonly used up to frequencies of 400 Mc/s.

Crystal diode mixers, developed during the war primarily for radar applications, are constructed in ceramic tubes using a silicon crystal and a tungsten "cat's whisker." A typical design is depicted in Fig. 53 which illustrates (*a*) the size, and (*b*) the construction. Diode valves, such as the EA50 may be used as mixers, but they are slightly more noisy than the crystal units quite apart from the difficulty of supplying the heater through RF chokes, which is a practical disadvantage.

A crystal mixer circuit suitable for operation in the 300 Mc/s. region is shown in Fig. 54. The tuned circuit consists of a coaxial line, shorted at one end and capacity loaded at the other. A capacity loaded line of this type

is preferable to an ordinary quarter-wave open ended line on the score of physical size; for example the 300 Mc/s. circuits shown in Fig. 55 are only 4" long.

The circuit is tuned to signal frequency by adjusting the knurled screwed cap on the end of the line, easily discernible in the photograph. The aerial feeder and mixer crystal are tapped down the line as shown and the position adjusted to give the desired compromise between selectivity and signal-to-noise ratio. Local oscillator injection is achieved through a small

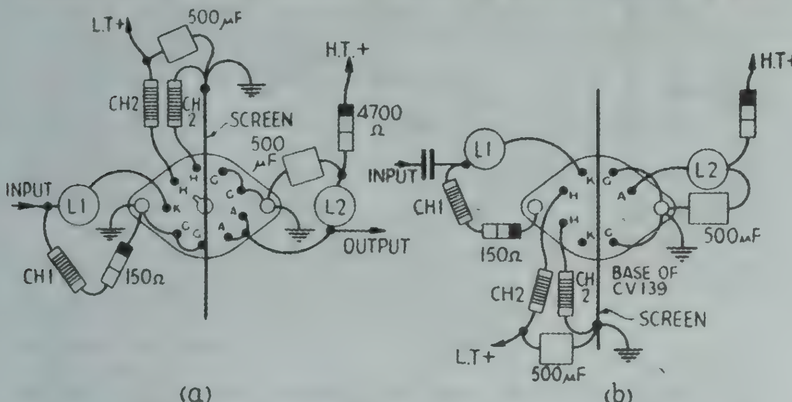


Fig. 52.
Layout of components in a grounded grid amplifier, (a) for a CV66 type valve and (b) for a CV139 type valve.

adjustable probe. The crystal current will contain DC, IF, signal frequency and local oscillator frequency components. The VHF components are by-passed to earth through the small capacity condenser C1 which is built into the crystal holder. The IF component is selected by the tuned circuit L2, C2 and passed on to the IF amplifier. A jack X in the earthy end of the IF circuit is used to take a meter which measures the DC crystal current.

Fig. 53.
Photographs of crystal diode mixer units showing (a) size and (b) a cut-away view showing the construction. (General Electric Co. Ltd.).

The latter may be set to about 0·25 mA. by adjustment of the local oscillator injection probe. Mixer crystals are easily damaged if too large a current is passed through them.

Another type of 300 Mc/s. tuned circuit which lends itself to experimental work, owing to its ease of construction and adjustment is shown in Fig. 56. This circuit consists of a copper, brass or aluminium trough of square cross

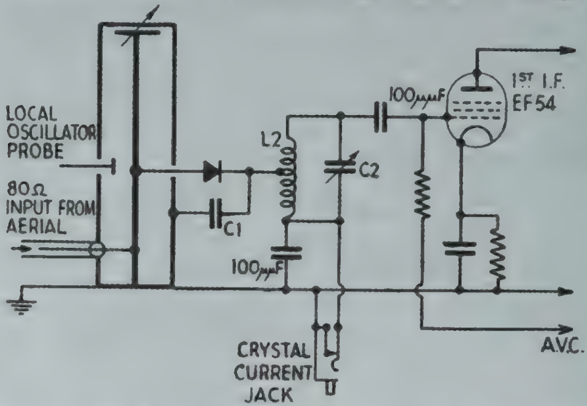


Fig. 54.
A crystal mixer circuit for 300 Mc/s. operation.

section closed at both ends containing an inner conductor fixed to one end of the trough. A parallel plate tuning condenser is used at the other end to facilitate tuning over a small band of frequencies. Aerial, local oscillator and mixer connections may be fed in through the bottom or sides of the trough as desired. For optimum *Q* the ratio of the length of the side of the square section of the trough to the diameter of the inner conductor, should be about 3·5 to 1.

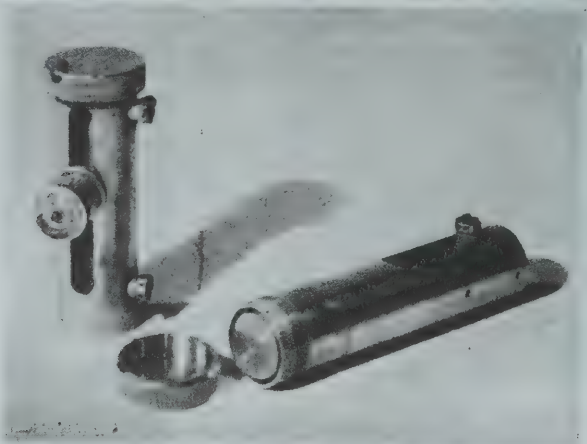


Fig. 55.
Capacity loaded line circuits suitable for 300 Mc/s. operation. (General Electric Co. Ltd.).

The following formulæ may be used to calculate the capacity required to tune a given length of line to a desired frequency:—

For Coaxial line circuits.

$$C = \frac{1}{138 \log \frac{b}{a} \left(\tan \frac{360 \times L}{\lambda} \right)} \times 10^{12} \ \mu\mu F.$$

where C is the capacity required in $\mu\mu$F.
b is the inside radius of the outer conductor.
a is the outside radius of the inner conductor in the same units as b.
L is the length of the line.
λ is the wavelength in same units as L.

For Open Parallel wire lines.

$$C = \frac{1}{276 \log \frac{D}{r} \left(\tan \frac{360 \times L}{\lambda} \right)} \times 10^{12} \ \mu\mu F.$$

where symbols are as above and:
D is the centre to centre spacing of the conductors.
r is the radius of each conductor in the same units as D.

For square trough type circuits the coaxial line formula may be used to give an approximate result, using b as the length of one side of the square section of the trough. Fig. 57 illustrates these formulæ.

(g) Local Oscillators

There are several valves which are suitable for use in the local oscillator section of a VHF superheterodyne receiver. As with RF and mixer stages there is a change in oscillator technique from lumped circuits to line circuits in the region of 200 Mc/s. Table V gives a list of common British and American valves suitable for use as local oscillators in VHF receivers.

TABLE V.

Valves suitable for use as Local Oscillators in VHF Receivers.

Valve.	Type.	Base.	Upper Frequency Limit (approx.) Mc/s.	Maker.
EF50	Pentode	B9G	200	Mullard
Z90	Pentode	B9G	200	Osram
DET20 ..	Triode	Octal	350	Osram
CV6	Triode	Octal	350	Service Type
Z77	Pentode	B7G	250	Osram
8D3	Pentode	B7G	250	Brimar
CV138 ..	Pentode	B7G	250	Service Type
9002	Triode	B7G	400	American
6C4	Triode	B7G	300	American
HA2	Triode	Acorn	600	Osram
955	Triode	Acorn	600	American
HY615* ..	Triode	Octal	350	American
EC52	Triode	B9G	400	Mullard

* Similar to Osram DET20.

The same general remarks, about short leads and efficient by-passing, apply just as much to the local oscillator as they do to the RF and mixer stages of a receiver. For trouble-free operation particular care should be taken with the mechanical design and construction of the oscillator section. A shoddy arrangement makes for critical erratic tuning and excessive frequency drift. Components should be chosen with care, and ceramic

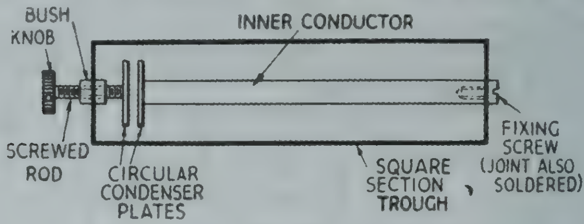


Fig. 56.
Plan sketch of a trough type capacity loaded line circuit.

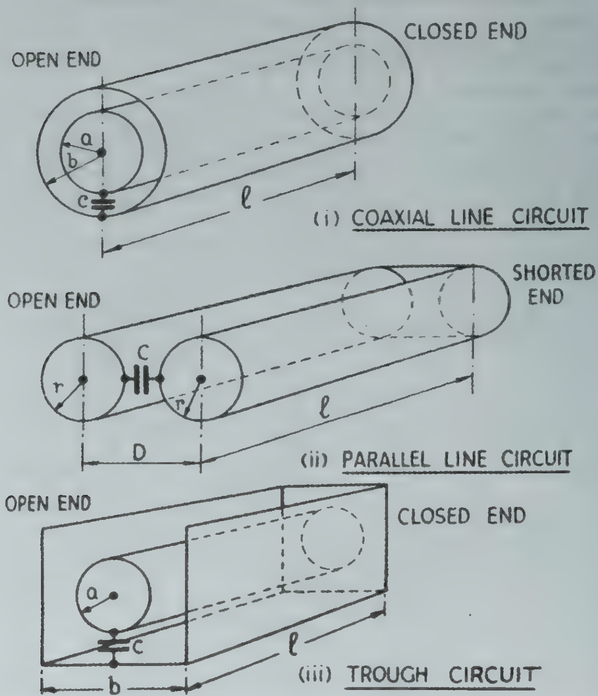


Fig. 57.
Common types of line circuit to illustrate the formulæ given on page 55.

Fig. 58.
The EC52 VHF triode oscillator valve suitable for operation up to 400 Mc/s. (Mullard Wireless Service Co. Ltd.).

condensers of the cup, disc or tube types should not be used across the tuned circuit or as grid condensers on account of their large temperature coefficients. Silvered mica or air spaced condensers are to be preferred.

Several local oscillator circuits, suitable for use up to about 200 Mc/s. are shown in Fig. 59, using conventional L and C circuit technique. Such coil and condenser circuits may be used in receivers operating well above 200 Mc/s. by injecting one-half or one-third of the required frequency into the mixer stage. The optimum conversion gain of a steep-slope pentode mixer stage will be found to be reduced by approximately one-half or one-third for one-half and one-third frequency injection respectively. Harmonic mixing in this way can be used with advantage to produce a receiver with a greater frequency stability and freedom from pulling. The following figures serve as an indication of the change in conversion gain of a CV138 mixer stage with harmonic order.

These results were obtained with a mixer circuit of the type shown in Fig. 60.

For high stability, the oscillator tuning coils should be wound on small ceramic formers, suitable examples of which are offered by several dealers and manufacturers. In order to reduce the effects of variation of valve capacities with temperature and supply voltage, the oscillator valve should be " tapped

Signal input for standard output (at 100 Mc/s.).	Local oscillator frequency (IF 10 Mc/s.).	Injection voltage for optimum conversion gain.
1 μV	90 Mc/s.	4 V
2 μV	45 Mc/s.	7 V
3 μV	30 Mc/s.	15 V

down " on the tuned circuit as far as possible consistent with the continuance of reliable oscillation. This tapping down may be performed directly on the coil or by the use of small grid and anode coupling condensers. The second method is probably the simplest in practice.

The local oscillator in a VHF superheterodyne should not be run too hard. The amplitude of oscillation need only be sufficient to provide the required mixer injection voltage. Too strong a local oscillator may give rise to spurious responses, and for this same reason a measure of screening round the oscillator stage is of advantage. Sometimes the same station may be heard at several points on the tuning dial, or the local oscillator may produce a rushing

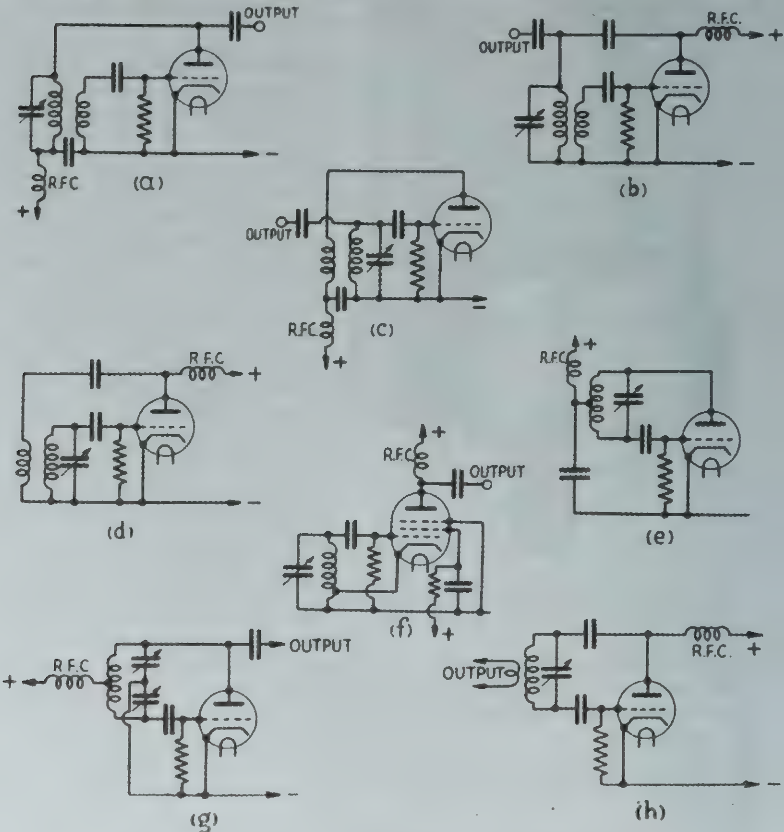


Fig. 59.
Local oscillator circuits suitable for use up to 200 Mc/s.

(a) Series fed tuned anode circuit.
(b) Shunt fed tuned anode circuit.
(c) Series fed tuned grid circuit.
(d) Shunt fed tuned grid circuit
(e) Series fed Hartley circuit.
(f) Electron coupled oscillator.
(g) Colpitts oscillator circuit.
(h) Ultraaudin oscillator circuit

sound when picked up on another receiver. These troubles may be due to "squeggering" of the local oscillator which can be cured by decreasing the value of the oscillator grid leak.

For mobile and point-to-point, operation, where fixed frequency operation is required, it is common practice to use receivers with crystal controlled local oscillators. A typical circuit is given in Fig. 60. For operation on 100 Mc/s. with an IF of 10 Mc/s. the crystal frequency used was 7·5 Mc/s. with a frequency multiplication of twelve—four times in the anode of the crystal oscillator and three in the following tripler stage. Crystal controlled receivers of this type are very useful for systems set-up to study propagation or aerial problems, as frequency drift is small, eliminating the need for retuning from time to time. The coupling circuit* between the RF and mixer stages has some advantages at VHF.

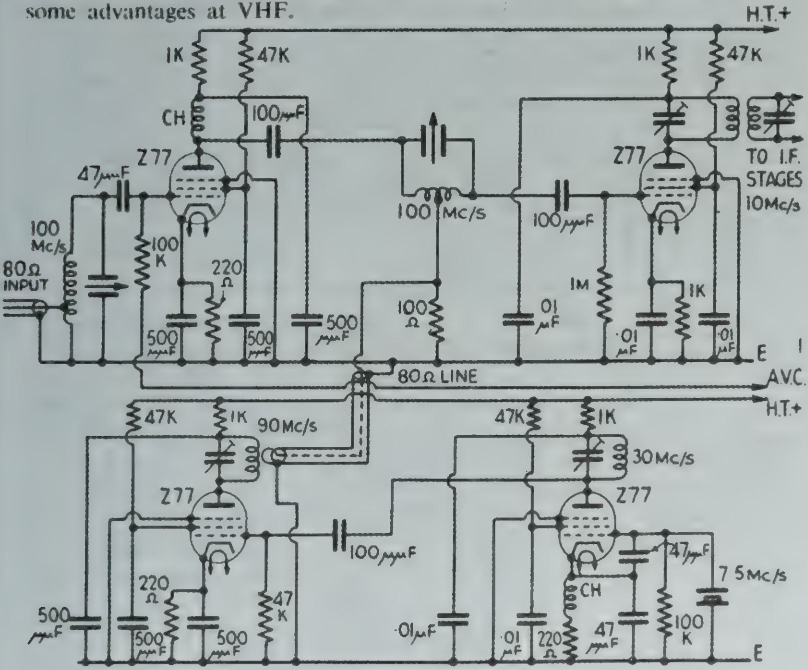


Fig. 60.
Circuit of RF and mixer stages with crystal controlled local oscillator.

Quarter-wave line type circuits may be used with advantage as tank circuits for VHF receiver local oscillators at frequencies above about 150 Mc/s. Below this frequency, line circuits are rather large and cumbersome.

Several types of line oscillator circuits are shown in Fig. 61. The circuits fall broadly into two groups, namely those using parallel line circuits and those using coaxial circuits. As the frequency is increased up to the region of 300 Mc/s. or higher, the coaxial circuit is preferable as it has lower losses. The trough type of construction of coaxial lines, already mentioned on page 54, may be applied to oscillator construction.

* *G. E. Co., Ltd.* Prov. Pat. No. 4651/45.

For maximum stability the line circuits should have the maximum practical Q, and in order to preserve a high figure the valve should be tapped well down the line. In the case of coaxial lines the best value of Q is achieved by making the ratio of the outer to inner conductor diameter equal to 3½ to 1.

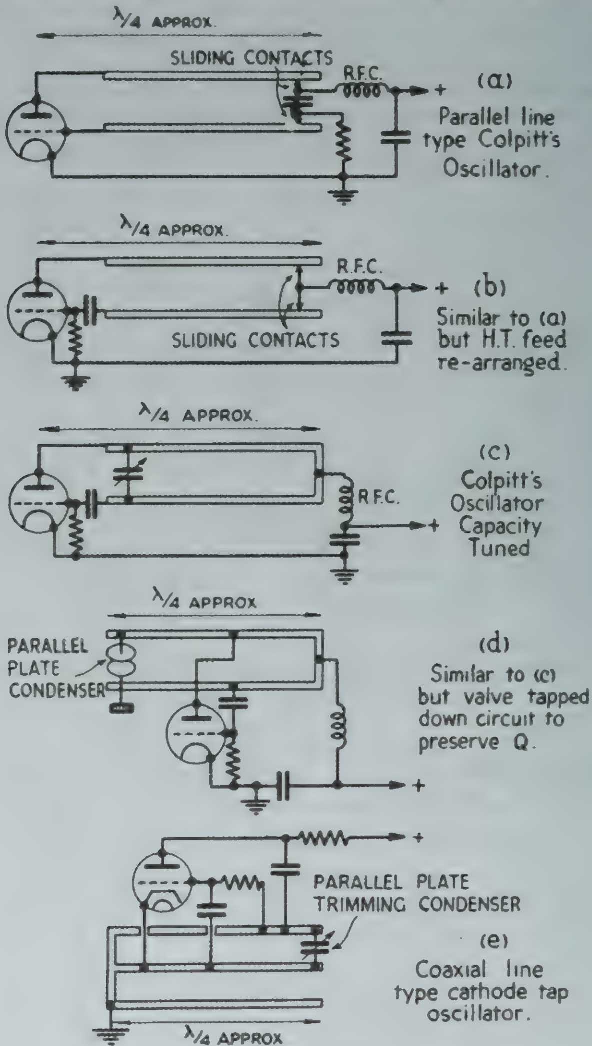


Fig. 61.
Single valve VHF line type oscillator circuits.

Owing to the fact that the valve and tuning capacities are placed across the circuits, the mechanical length will be considerably less than an electrical quarter wavelength. The discrepancy will be less when the valve is tapped down than when it is placed across the open end of the quarter-wave line. The formulæ on page 55 may be used to calculate the approximate length of the line required if the valve interelectrode capacities are known. The tuning capacities across lines should be kept as small as possible in order to preserve the Q of the circuit as a whole. A coupling loop or a probe may be used to couple the line oscillator to the mixer valve of a receiver.

Another VHF oscillator of high stability is the "pot" type shown in Fig. 62. The "pot" tank circuit has high Q and a low L to C ratio. By suitably choosing the materials from which the circuit is constructed a high degree of stability with temperature change may be achieved.

Exact dimensions for a given frequency are a little difficult to calculate, but the following dimensions taken from the *Radio Amateurs' Handbook*, published by the A.R.R.L., are given as a guide:—

Frequency Mc/s.	Outer Cylinder.	Inner Cylinder.	Inner Rod.
112	3″ dia. × 3¾″	2½″ dia. × 2¾″	¾″ dia. × 3½″
224	3″ dia. × 3″	2½″ dia. × 2½″	¾″ dia. × 2¾″
400	2″ dia. × 3″	1½″ dia. × 1½″	½″ dia. × 1¾″

(h) Beat Frequency Oscillators

The reception of CW signals is possible using crystal controlled transmitters and stable receivers up to a frequency of about 150 Mc/s. Above this frequency CW reception is not practical as the beat note is likely to be too unstable and is liable to drift beyond the limit of audibility.

For CW reception a beat frequency oscillator is required as in normal superheterodyne practice. The BFO may consist of an electron coupled oscillator operating at the receiver intermediate frequency, feeding into the second detector. The oscillator should be carefully screened and oscillations

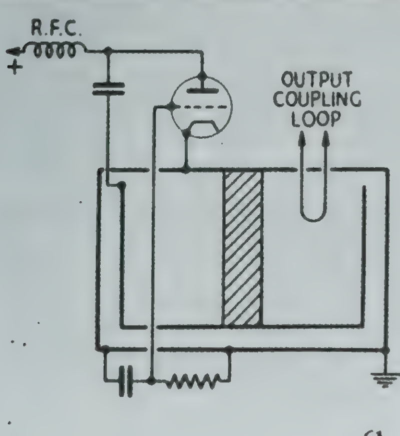


Fig. 62.
"Pot" type oscillator circuit.

should be as weak as possible consistent with a loud enough beat note. The L to C ratio of the oscillator tuned circuit should be low in order to minimise the strength of harmonics which may extend up to the signal frequency region and give rise to spurious whistles.

(i) Automatic Volume Control

Automatic volume control is desirable in a VHF superheterodyne receiver where signals at levels between one microvolt and several hundreds of millivolts are commonplace. Fading, due to multi-path transmission, can occur at VHF, particularly when transmission is being made over distances in excess of the horizon distance.

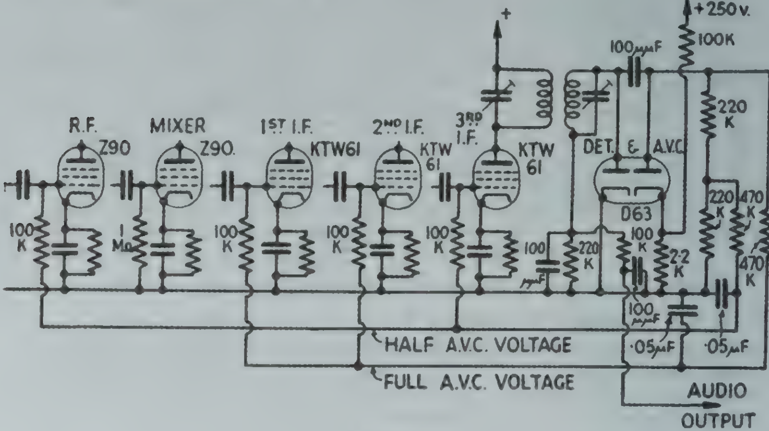


Fig. 63.
Circuit of a delayed Automatic Volume Control system.

AVC technique in VHF receivers follows normal AVC practice. A portion of the AVC voltage may be applied quite safely to sharp cut-off RF amplifier stages where the signal is small and, therefore, little distortion will occur. Full AVC voltage may be applied to the IF stages if variable-*mu* valves are used, but it is not recommended that AVC should be applied to mixer valves. This is for two reasons. First, because change of bias voltage on the mixer may alter the local oscillator frequency, and second, because if the local oscillator injection is sufficient to cause grid current to flow in the mixer circuit then the other stages would become biased back thereby causing a decrease in receiver sensitivity.

Simple delayed AVC is recommended for amateur receivers, using a delay voltage of some 5 to 10 volts. A suitable circuit is shown in Fig. 63.

In the interests of preserving a high signal-to-noise ratio, in an elaborate set, two AVC supplies may be used with different delays. The AVC supply with the larger delay should be used to supply the RF stage and the other for the IF stages. In this way the IF gain is the first to be decreased as the signal level is increased leaving the RF stage, or stages, at full amplification. The AVC is then allowed to operate on the RF section when the signal level is high compared with the receiver noise level.

Amplified AVC may be used if an exceptionally flat AVC characteristic is desired. A suitable circuit is given in Fig. 64. The circuit operates as follows. The rectified carrier voltage is smoothed and applied between grid and cathode of the triode DC amplifier valve. In the absence of a signal this voltage is zero and the cathode potential of the triode is positive with respect to earth, depending on the value of the resistance R1. In this condition the diode is non-conducting and the AVC line is biased to a steady −2 volts. As the signal increases, the grid potential of the triode, with respect to the cathode, becomes negative and the cathode potential of the triode first falls to zero and then becomes negative with respect to earth. The delay diode will then conduct and an amplified AVC voltage will be applied to the AVC line. The delay voltage may be adjusted for the desired carrier level at the detector by the choice of a suitable value for R1.

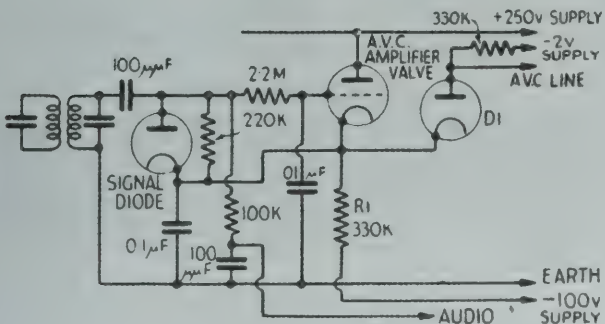


Fig. 64.
Circuit of an amplified Automatic Volume Control system.

The Double Superheterodyne Receiver

For several reasons the double superheterodyne principle is very attractive in VHF practice.

(1) A high first IF enables a high degree of second channel rejection to be achieved.

(2) By splitting-up the receiver gain into sections at signal frequency, first IF and second IF, it is easy to achieve a high overall gain with stability.

(3) The use of a low second IF provides a high degree of adjacent channel selectivity.

(4) A variable frequency second local oscillator allows for band spreading over a limited frequency range (governed by the selectivity of the RF and first IF circuits). The degree of bandspreading is the same whatever the setting of the main tuning control, *i.e.* the same number of kilocycles per division are accommodated on the bandspread dial at all settings of the main tuning control.

The most common disadvantage of the double superheterodyne is the presence of spurious responses. By enclosing the second local oscillator and mixer, together with its input and output IF transformers, in a screening box, and by employing filtered HT, LT and AVC leads these spurious responses may be reduced to negligible proportions.

Fig. 65 shows a block diagram of a double superheterodyne receiver suitable for operation up to about 150 Mc/s. As an example the set is shown receiving a 60 Mc/s. signal. The incoming signal is passed through an RF amplifier into a first mixer where it is mixed with a 50 Mc/s. local oscillation to produce a first IF of 10 Mc/s. A first IF of 10 Mc/s. in this example gives a second channel response 20 Mc/s. below the signal frequency, *i.e.* at 40 Mc/s., which

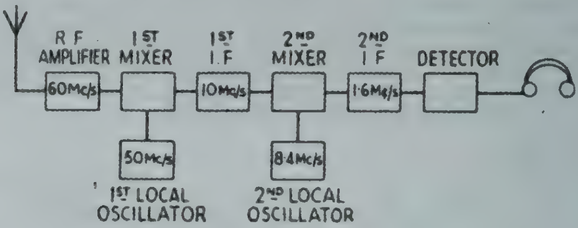


Fig. 65.
Block diagram of a double superheterodyne receiver.

is easily rejected by two RF tuned circuits. The 10 Mc/s. IF signal is passed through an amplifier and then into a second mixer, where it is mixed with an 8·4 Mc/s. local oscillation to give a 1·6 Mc/s. second IF which passes on to a detector and LF stages as in ordinary superheterodyne practice. A frequency of 1·6 Mc/s. was chosen for the second IF as being the lowest which will give a suitably wide bandwidth (about 20 kc/s.) with simple coils critically coupled.

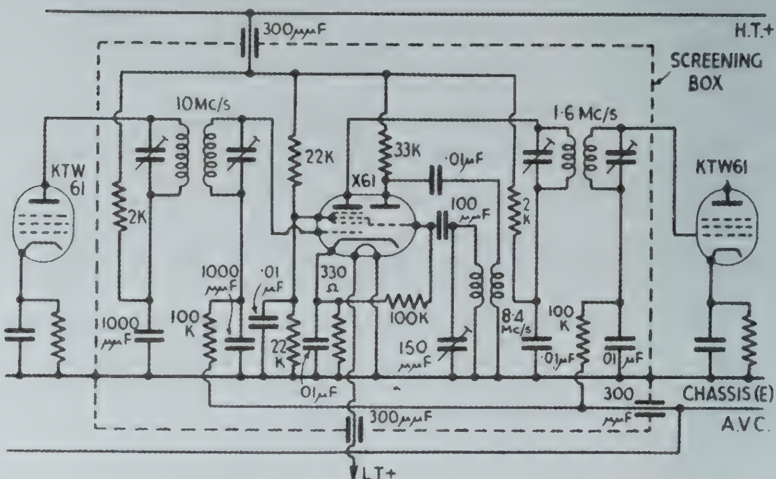


Fig. 66.
Circuit diagram of a VHF double superheterodyne second frequency changer showing the screening necessary to avoid spurious whistles.

Fig. 66 shows the second frequency changer and associated circuits of a double superheterodyne receiver indicating clearly the section which should be totally screened in a box in order to eliminate spurious responses. The feed-through condensers shown in Fig. 35 are very suitable for filtering the leads entering the screening box.

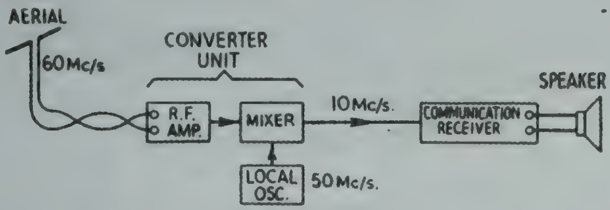


Fig. 67.
Block diagram of a VHF superheterodyne converter system.

Superheterodyne Converters

A superheterodyne converter consists essentially of a frequency changer stage (with or without an RF stage) used in front of an ordinary communications receiver to extend its frequency coverage. The communications receiver is tuned to a fixed frequency, say 10 Mc/s., at which frequency it acts as an IF amplifier and audio section of the composite receiver. The converter unit provides an IF output (10 Mc/s. in this case) which is fed into the communications receiver input terminals. This is illustrated in block diagram form in Fig. 67.

Superheterodyne converters may be used up to 60 Mc/s. but above that figure the bandwidth of the average communications receiver is too narrow to enable signals to be held in tune without some form of automatic frequency control consequently tuning becomes too critical.

The circuits shown in Figs. 51 and 48 with a suitable local oscillator provided may be used as VHF converters. A coupling link wound round the earthy end of the IF transformer primary may be used to couple *via* a coaxial lead to the input terminals of the communications receiver.

CHAPTER 5 VHF MEASUREMENTS

FREQUENCY MEASUREMENT—TRANSMITTER POWER OUTPUT MEASUREMENT—RECEIVER PERFORMANCE MEASUREMENT.

THERE are a few important but simple measurements which the amateur may make to ensure that his apparatus is operating correctly. These measurements include:—

(*a*) Frequency measurement.
(*b*) Transmitter power output measurement.
(*c*) Receiver performance measurements.

Frequency Measurement

Above about 100 Mc/s. it is practicable to measure directly the wavelength at which a transmitter or oscillator is operating by using Lecher wires. Lecher wires consist of a pair of taut parallel wires, spaced an inch or so apart to form an open wire transmission line, and a bridge to short circuit the wires. The latter device can be slid along the line as required. Fig. 68 shows a convenient method of constructing such a line.

For transmitter wavelength measurement one end of the line is loosely coupled by a loop, to the tank circuit of the transmitter, or oscillator, whose wavelength is required. Starting near the coupling loop end of the line, the bridge should be slowly moved along towards the open end of the line until a point of maximum current in the bridge is found; this will be indicated by a deflection on the transmitter anode current meter, or by noticing when a pea lamp and loop loosely coupled to the tank coil passes through a minimum in brightness. This position should be carefully noted and the bridge then

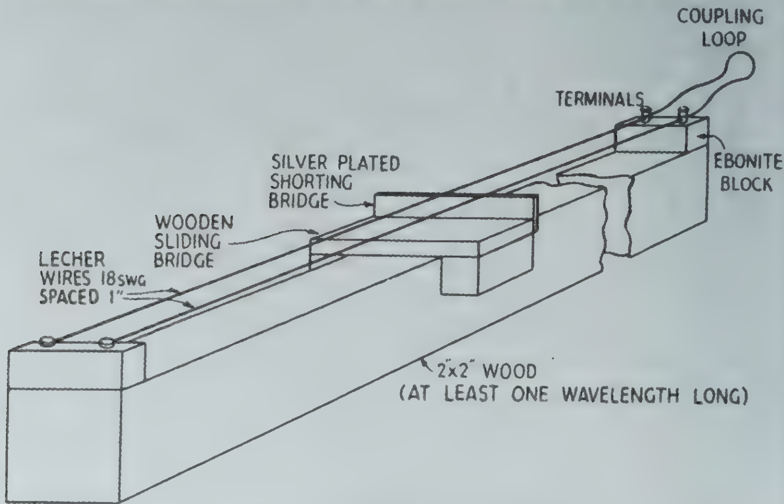


Fig. 68.
Construction of a simple Lecher wire system.

moved further along the line until the next consecutive similar position is found. The distance between the two points will be one-half of the wavelength at which the transmitter is oscillating. The experimental set-up is shown in Fig. 69.

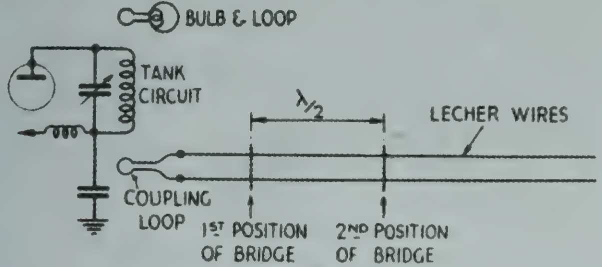


Fig. 69.
Measuring wavelength using Lecher wires.

An application of the formula,

$$Frequency = \frac{15{,}000}{Distance\ between\ bridge\ positions\ in\ cm.}\ Mc/s.$$

will enable the frequency of the oscillation to be determined. For example if the distance between the bridge positions is found to be 100 cm., then by substituting in the formula,

$$Frequency = \frac{15{,}000}{100}\ Mc/s. = 150\ Mc/s.$$

For the most sensitive condition of adjustment, the Lecher wires and the bulb and loop should be very loosely coupled to the tank circuit. This is especially important when measuring the frequency of a self-excited oscillator whose frequency may be altered if the coupling is too tight. Although the accuracy of frequency measurement by this method is not very high the method is of great use to the amateur as the only instrument required is a good ruler. An accuracy of $0 \cdot 1\%$ can be attained with care.

In the case of receiver frequency measurement, with straight sets and super-regenerative sets, the line should be loosely coupled to the detector tuned circuit, with the receiver reaction control adjusted so that oscillation is just taking place in the case of the straight set, or the characteristic hiss is just present in the case of the super-regenerative set. Sliding the bridge along the line, as for transmitter measurement, the current maxima in the bridge will be indicated by noting when oscillation or super-regeneration ceases. The coupling to the receiver tuned circuit should be reduced until the oscillations are only just stopped as the bridge is moved through the position of the

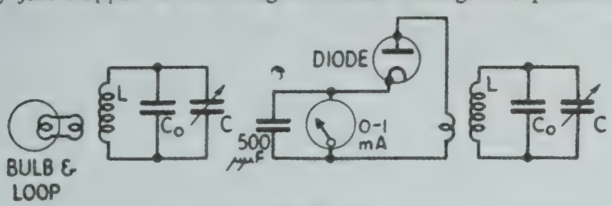


Fig. 70.
Two absorption wavemeter circuits.

current maxima. As in the transmitter case the distance between the bridge positions is one-half of the wavelength to which the receiver is tuned. The corresponding frequency may be calculated as already explained.

In the case of superheterodyne receiver frequency measurement the line may be coupled to the local oscillator tuned circuit. The positions of the current maxima as the bridge is moved along the line are most conveniently found by noting when the reading on a grid current meter, inserted in the earthy end of the oscillator grid leak, passes through a minimum. The coupling should only be just tight enough to enable the deflection on the grid current meter to be discernible. Having found the frequency at which the local oscillator of the receiver is operating the corresponding signal frequency may be found by adding or subtracting the receiver intermediate frequency, depending on whether the local oscillator is designed to operate below or above the signal frequency.

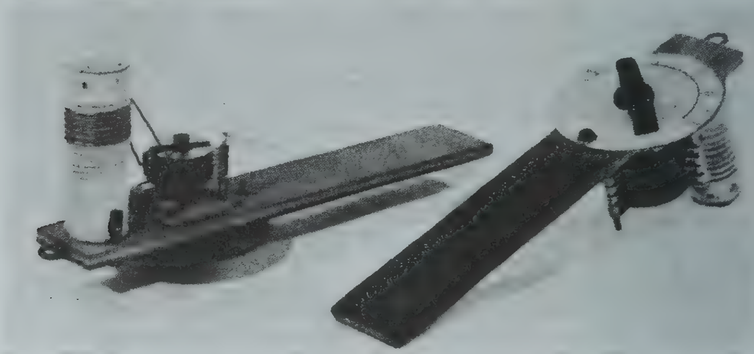


Fig. 71.
Photograph of two absorption wavemeters covering 20 to 60 Mc/s.

Absorption wavemeters are very convenient types of measuring instrument for rough frequency checks, *e.g.* for locating the amateur bands. They consist of a simple tuned circuit of high Q and good stability with some device to indicate resonance. The resonance indicator usually takes the form of either a bulb and loop or a diode rectifier and microammeter. In each case the resonance indicator should be as loosely coupled to the tuned circuit as possible consistent with giving a suitable indication in order that the high Q of the circuit shall be preserved. Two such wavemeter circuits are shown in Fig. 70. Fig. 71 illustrates one convenient method of constructing an absorption wavemeter. The coils shown are tension-wound on ceramic formers in order to give a high degree of electrical and mechanical stability. The two wavemeters shown cover the range 20–40 Mc/s. and 40–60 Mc/s.

Another interesting device is shown in Fig. 72. This type has been called the "Barrel" wavemeter on account of its shape. The paxolin body contains an *Eddystone* 100 $\mu\mu$F variable condenser and the inductance is made up of a G-shaped piece of silver-plated brass easily seen in the photograph. The variable condenser bush is fixed through a $\frac{3}{8}''$ diameter hole in the cross bar of the G. An arm with a phosphor bronze wiper is also fixed to the rotor. The wiper runs round the curved portion of the G as the condenser is rotated. The top end of the curved portion of the G is fixed to the stator plates of the tuning condenser. The whole assembly forms a variable inductance and

variable capacity circuit of high Q. A pea lamp connected across a portion of the rotating arm may be used as a resonance indicator. The instrument shown has a range of from 56–330 Mc/s. Details of the size of the G-shaped piece of brass are given in Fig. 73.

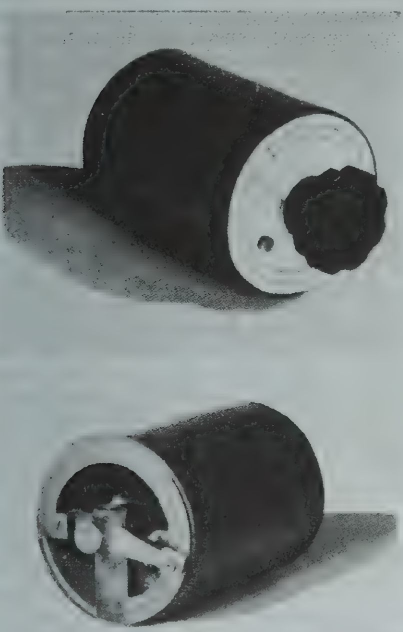


Fig. 72.
The "Barrel" type wavemeter covering 56 to 330 Mc/s. in one range.

Absorption wavemeters of the types just described may be calibrated using a self excited oscillator which has been previously calibrated by the Lecher wire method or some other means. The wavemeter should be held near the oscillator tank coil and tuned until the oscillator grid current meter reading falls to a minimum or until the resonance indicator on the wavemeter shows a maximum indication. The wavemeter should then be moved gradually further away from the tank coil until the indications mentioned above are only just discernible; during this operation the wavemeter tuning should be continuously adjusted for resonance. A mark may then be made on the wavemeter scale corresponding to the wavelength at which the oscillator is operating.

Heterodyne wavemeters are very useful for checking receiver frequencies. A simple type circuit is shown in Fig. 74. Great care should be taken with the construction of the tuned circuit in order to achieve a high degree of electrical and mechanical stability. The meter shown in Fig. 74 has already been described in some detail in *The Amateur Radio Handbook*, to which the reader is referred for further details. Calibration may be accomplished by using Lecher wires, or if higher accuracy is required by comparing the wavemeter with known harmonics of a crystal oscillator. The Lecher wires may be helpful in deciding which crystal harmonic is being used.

A more ambitious type of heterodyne wavemeter may employ its own quartz crystal as a harmonic generator. Harmonics of a 1 Mc/s. crystal-controlled oscillator may be heard up to 150 Mc/s. The circuit of Fig. 75 shows a 1 Mc/s. crystal oscillator driving a harmonic generator valve. A further refinement would be the provision of a 10 Mc/s. crystal oscillator giving reference harmonics 10 Mc/s. apart. The heterodyne wavemeter may be checked by plugging a pair of headphones into the jack and tuning the oscillator for zero beat with one of the crystal harmonics. It is best to check against the 10 Mc/s. harmonics first and then substitute the 1 Mc/s. harmonics, otherwise there may be some ambiguity as to the

particular harmonic against which the scale of the meter is being checked. When tuned to zero beat with the desired harmonic, the pointer should exactly coincide with the scale calibration corresponding to the crystal harmonic. If the pointer is arranged so that it can be moved relative to the scale, but leaving the tuning condenser in the same position, then any errors in the calibration can be taken up by adjustment of the pointer.

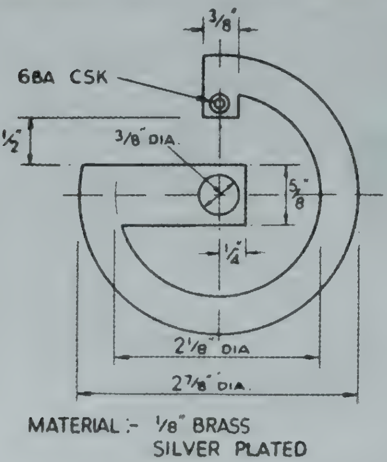


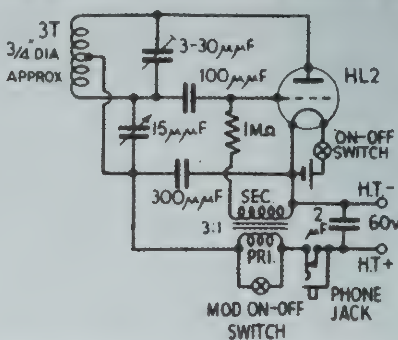
Fig. 73.
Constructional details of the "Barrel" wavemeter inductance element.

To measure the frequency of a transmitter, the calibration of the wavemeter should first be checked (near the frequency at which the transmitter is operating) against the crystal calibrator. Tuning the frequency meter for zero beat in the headphones, with the crystal oscillator switched off, the transmitter frequency may be read-off directly from the wavemeter scale. Measurements should not be made with the crystal oscillator operating, as many spurious whistles may be introduced. For received signal frequency measurement, after checking the wavemeter calibration against the crystal harmonics, the wavemeter should be set so as to give zero beat with the incoming signal of the receiver. A similar method may be used to calibrate a receiver, adjusting the wavemeter for zero beat with the receiver BFO (if it has one), or adjusting the meter to give a maximum reading on the tuning indicator of the set.

Transmitter Power Output Measurement

Transmitter power output measurements at VHF are important to enable the operator to ascertain if the output stage of his transmitter is operating so

Fig. 74.
A simple heterodyne wavemeter circuit.



as to give the maximum possible output of RF energy without exceeding the rated anode dissipation of the valve. The measurement can conveniently be carried out by using the RF power to light a load lamp, and comparing the brightness of the lamp with that of a similar lamp operated from a known DC or low frequency AC supply.

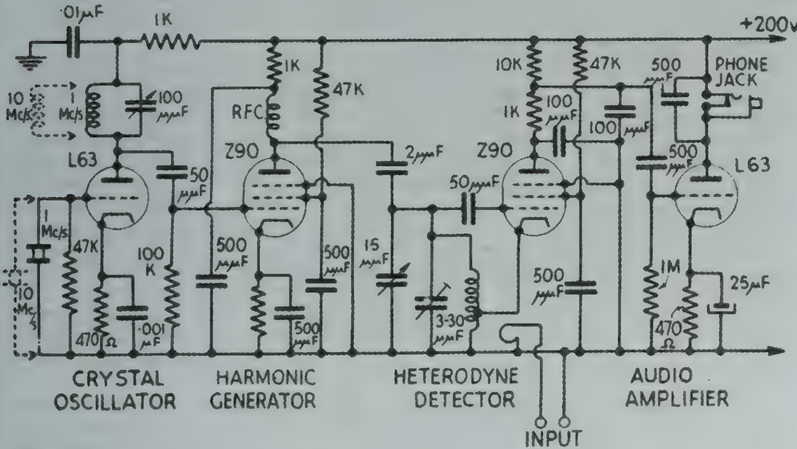


Fig. 75.
A heterodyne wavemeter circuit with quartz crystal checking unit.

Fig. 76 shows the experimental set-up required. L1, C1 is the transmitter tank circuit to which the load lamp is coupled by means of a one or two turn link coil L2, in series with a 50 $\mu\mu$F variable condenser C2. The load circuit should first be tuned to resonance with the tank circuit, using very loose coupling, by adjusting C2 and noting when the load lamp just glows red. The coupling between the circuits should then be increased until the output stage is drawing the correct anode current. Some slight retuning of the load or

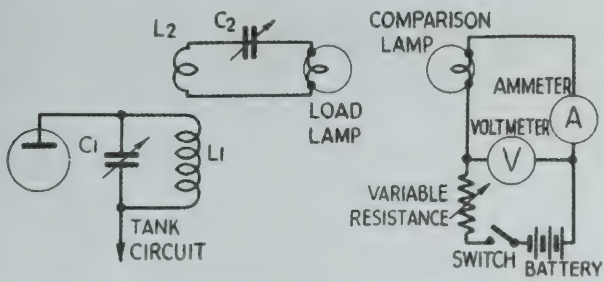


Fig. 76.
Experimental set up for the measurement of transmitter power output.

tank circuit may give improved results. The comparison lamp of the same make and power as the load lamp may next be switched on and its brilliance adjusted by means of the variable resistance until it is the same brilliance as the load lamp. The product of the current and the voltage read from the two meters will give the power in the comparison lamp which may be taken as being equal to the power output from the transmitter since the two lamps have been adjusted to equal brilliance.

Receiver Performance Measurements

It has been customary to express the performance of a VHF receiver by stating that the sensitivity is so many microvolts input to the aerial terminals for a certain signal-to-noise ratio—usually 20 db. The measurement is carried out in the following way. A signal generator is coupled through a suitable matching resistance to the input terminals of the receiver under test. The generator then tuned to the receiver frequency and the attenuator turned down until a weak signal is being fed into the set. A note of the noise output from the set should be made with no modulation using an output meter or oscilloscope as an indicator. The modulation on the signal generator should then be switched on at a level of 30%. The new reading on the output meter should be noted and compared with that prevailing with no modulation. Repeating the test at different carrier input levels will enable the input level which gives a 20 db signal-to-noise ratio to be found. In a good VHF communication receiver this figure should be a few microvolts.

In the absence of a signal generator the following test may be applied to see if a receiver is of good quality. The aerial circuit should be tuned through resonance while the noise output of the set is observed on some form of output meter. In a good receiver the noise voltage will vary 2 to 1 or more as this test is carried out.

CHAPTER 6 FREQUENCY MODULATION

CW TELEGRAPHY—AMPLITUDE MODULATION—FREQUENCY MODULATION—PHASE MODULATION—ADVANTAGES OF FM—DISADVANTAGES OF FM—REACTANCE VALVE MODULATION—ARMSTRONG'S PHASE MODULATOR—PRACTICAL FM EXCITER UNIT—FM RECEIVER CONSIDERATIONS—LIMITERS—RECEIVER DESIGN (GENERAL)—NOISE REDUCTION IN FM SYSTEMS—PRE-EMPHASIS—MEASUREMENT OF FREQUENCY DEVIATION.

AT first sight the system of transmission known as Frequency Modulation may appear to call for the use of complicated and expensive apparatus, difficult to adjust outside a laboratory. In point of fact, however, this is not the case, for as will be shown later, it is a comparatively simple matter to construct a FM transmitter and receiver using components of a type normally available to the amateur.

The first experiments with FM were carried out nearly 20 years ago, which disposes of the myth that it is a new invention. As FM is only one of several methods which are available of conveying intelligence by means of radio waves, it is proposed to consider briefly the systems in use, with a view to preparing the way for a detailed description of FM.

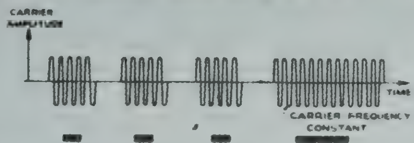


Fig. 77.
Pictorial representation of the letter " V " using the Morse code.

Continuous Wave Telegraphy

In this system a carrier wave at radio frequency is interrupted in a predetermined manner by means of a key; different combinations of dots and dashes representing letters and numerals of the Morse Code. A pictorial representation of the Morse letter V is given in Fig. 77.

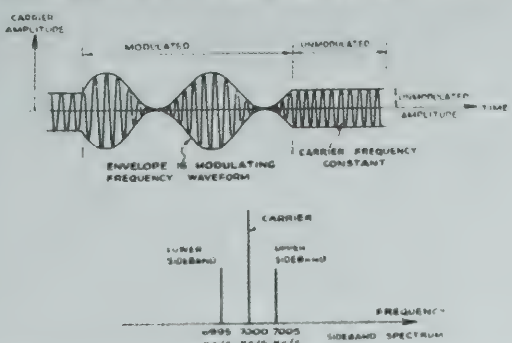


Fig. 78.
Amplitude modulated wave and sideband spectrum of example referred to in text.

For the transmission of speech and music a carrier wave at radio frequency is again employed, but instead of interrupting it, as for the CW case, its amplitude, frequency or phase is varied in sympathy with audio frequency speech or music waves, thus giving rise to Amplitude, Frequency or Phase Modulation.

Amplitude Modulation

In this system a constant frequency carrier wave (at radio frequency) is used, its amplitude being varied, or modulated by the low frequency voltages produced at the microphone after suitable amplification. The change in amplitude of the carrier is made proportional to the instantaneous audio voltage, and the rate at which the amplitude is varied is equal to the frequency of the audio voltage.

It can be shown mathematically that an AM wave consists of a carrier frequency and a set of side-bands. A 7,000 kc/s. carrier, for example, when amplitude modulated by a single 5 kc/s. audio note, is equivalent to (1) a carrier at 7,000 kc/s., (2) an upper side-band at 7,005 kc/s., and (3) a lower side-band at 6,995 kc/s. Such a transmission thus takes up a total band-width of 10 kc/s. Fig. 78 shows the shape of an amplitude modulated wave and the positions occupied by the side-bands.

In the case of complicated modulating wave forms the side-band *spectrum* becomes very complex, although the side-bands themselves do not extend beyond the highest modulating frequency on either side of the carrier frequency.

Frequency Modulation

After the existence of side-bands had been established, in the AM case, it was anticipated that, by varying the carrier frequency in sympathy with the audio waves, the band-width required for the transmission of speech and

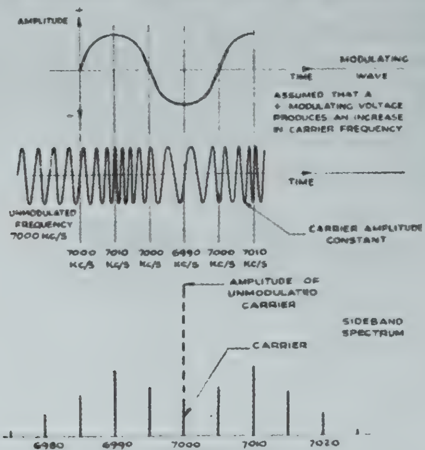


Fig. 79.

Frequency modulated wave and side-band spectrum of example referred to in text, showing that the spacing between the side-bands is equal to the audio frequency, namely 5 kc/s.

music could be reduced. For example it was thought that by using a deviation of ± 500 c.p.s., a band-width of only 1,000 c/s (1 kc/s.) would be required to transmit any audio frequency. In 1922, however, Carson showed mathematically that a frequency modulated wave was equivalent to a carrier plus an infinite number of side-bands, although in practice it is only the first few side-bands that are large, the remainder being negligible.

In the case of FM there is a carrier wave at radio frequency whose deviation from the mean carrier frequency is made proportional to the instantaneous audio voltage. The rate at which the carrier frequency is deviated is equal to the audio modulating frequency.

Consider a numerical example. Suppose a 7,000 kc/s. carrier frequency is modulated by a pure 5 kc/s. tone to produce a peak deviation of ± 10 kc/s, then during 100% modulation the carrier will be deviated between 7,010 kc/s, and 6,990 kc/s. at a rate of 5 kc/s. The lower portion of Fig. 79 shows the side-band spectrum corresponding to this example, whilst the upper portion gives a pictorial representation of the frequency modulated wave.

The fact that a frequency modulated wave can be shown to consist of a carrier and an infinite set of side-bands, does not mean that in practice an infinitely wide band-width is required to transmit FM signals. So long as the *deviation ratio*, *i.e.* the ratio of the peak deviation to the modulating frequency, is made greater than unity, then the side-bands outside the peak deviation become small enough to be neglected. This means that modulating frequencies up to 5 kc/s. can be transmitted over a system having a pass band-width of ± 5 kc/s., without excessive distortion, although for a high quality service such as broadcasting, where a high signal-to-noise ratio is essential, a greater deviation ratio and a wider receiver band-width are required.

Phase Modulation

In this system the phase of the carrier wave is altered in sympathy with the audio modulating voltage. Phase Modulation is not, at present, used to any great extent as a means of communication but PM transmitters, with simple corrector networks in the LF stages, are used to produce frequency modulation. Such a scheme is used in the Armstrong system of FM transmission.

There is a great similarity between frequency and phase modulation, in fact a person listening to a fixed note on a FM receiver would be unable to distinguish whether the transmission was frequency or phase modulated.

The relationship between frequency and phase modulation is simple and is expressed as:—

$$\delta f = n\,\emptyset$$

where δf = frequency deviation in c/s.
n = modulating frequency in c/s.
$\emptyset$ = phase deviation in radians ($180^\circ = 3{\cdot}14$ radians).

Expressed in words this means that phase modulation is a form of frequency modulation in which the deviation is proportional to the modulating frequency.

Consider the case of a PM transmitter with a constant deviation of ± 5 radians ($5 \times 360/2\pi^\circ$), then if the modulating frequency were 1,000 c/s the equivalent frequency deviation would be ± 5 kc/s. If the audio frequency were raised to 3,000 c/s the equivalent frequency deviation would then be ± 15 kc/s.

Fig. 80 shows the graphical relationship between equivalent frequency deviation and modulation frequency, for a constant phase deviation of 5 radians.

Having briefly surveyed the different methods of modulation, we shall now proceed to discuss the advantages and disadvantages of the FM system.

Advantages of FM

(1) The great reduction of set and impulse noise which can be achieved.

The former includes circuit and valve noise emanating from the HF stage of the receiver. Impulse noise includes motor car ignition and other forms of man-made static as well as certain types of atmospherics. As noise reduction takes place in the receiver it will be considered more fully in a later section.

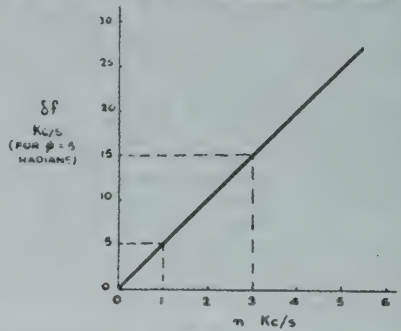


Fig. 80.

Graph showing relationship between equivalent frequency deviation and modulation frequency for a constant phase deviation of 5 radians.

(2) The saving of space and power in the transmitter.

In a frequency modulated transmitter, modulation is performed at low level in the master-oscillator stage, consequently a valve of the receiving RF pentode type is sufficient to modulate the largest transmitter. Expensive high power audio equipment and power supplies are thus dispensed with—the saving in cost being more marked as the transmitter power is increased.

(3) More ouput is available from a given type of valve.

Since the amplitude of the frequency modulated wave is constant, the output stage can be run under Class C telegraphic conditions. No voltage peaks occur during modulation as is the case with amplitude modulated transmissions. Thus for a given type of output valve more output power is available. Conversely for a given output power, smaller valves and tuning condensers are required than would be the case for an equivalent amplitude modulated transmitter. The net result of these economies is that a FM transmitter can be made approximately half the size of a high-level plate-modulated transmitter using AM and the same carrier power.

Disadvantages of FM

(1) The difficulty of keeping the mean frequency of the carrier constant in order not to interfere with transmitters working on neighbouring channels.

This disadvantage was gradually overcome as the technique developed. The difficulty of the problem becomes apparent when it is realised that the carrier frequency has to be capable of rapid frequency variation in order to transmit the intelligence, yet at the same time the *mean* carrier frequency must be held constant. The methods used to-day for obtaining this condition of stability are discussed later.

(2) The larger band-width required to provide a high signal-to-noise ratio service.

This disadvantage has been overcome by using and developing the frequency spectrum above 30 Mc/s. At these high frequencies many wide-band transmitters can operate without mutual interference.

In America, where a standard deviation of $\pm$ 75 kc/s. is allowed the FM broadcast band falls around 43 Mc/s., each station being allocated a channel-width of 200 kc/s.

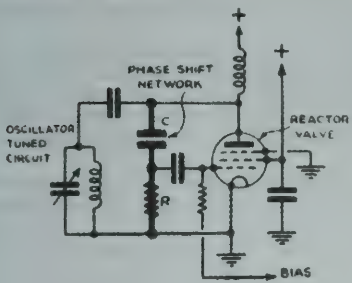


Fig. 81.

A typical reactance valve modulator showing the phase shift network in heavy lines.

Methods of Producing FM Transmissions

FM is no new idea, in fact, the old arc transmitters were in effect frequency modulated when the frequency was changed from "mark" to "space" during keying. Attempts were even made to frequency modulate arc transmitters with speech, but nothing came of the experiments.

With the introduction of the thermionic valve further attempts were made to modulate oscillators by methods such as that of connecting a condenser microphone across the tuned circuit of a self-excited oscillator. Again the experiments proved abortive and the doom of frequency modulation seemed to be sealed when Carson expounded his side-band theory in 1922.

During 1935 Major Armstrong, in the U.S.A., conducted some experiments, using frequency modulation, with a view to reducing noise on radio transmissions. A year later he published his paper in the *Proc. I.R.E.*, since when interest in FM both for broadcast and communication purposes has increased by leaps and bounds.

The most popular forms of FM transmitter at present in use are:—

(*a*) those which use the reactance valve modulator method, and

(*b*) those which use a phase modulator, with a correcting network to convert the phase modulation into frequency modulation.

Reactance Valve Modulators

The reactance valve (or reactor) was adapted for frequency modulation by Murray Crosby in America. In essentials it is a valve connected across an oscillator tuned circuit in such a way that it behaves as a variable capacity or inductance—the change in reactance being controlled by the grid voltage of the valve.

The fundamental point about a reactor is that its control grid is fed with an RF voltage 90° out of phase with the voltage across its anode. A typical reactor circuit is shown in Fig. 81, the resistance R and the condenser C forming the phase change circuit. The condition that the phase shift should

be 90° is that the reactance of C should be large compared with the resistance R. Fig. 82 shows a typical reactor characteristic—grid bias voltage being plotted against oscillator frequency. In practice the reactor is biased at A in the centre of the linear portion of the characteristic, and the modulating voltage is superimposed on this voltage. As the grid voltage varies between C and D, the frequency of the oscillator is caused to vary between E and F, the deviation being proportional to the amplitude of the audio voltage and the rate of frequency swing being equal to the modulating frequency. Thus pure frequency modulation is produced so long as the grid voltage excursion keeps to the straight part of the characteristic.

The single reactor type of FM transmitter is not altogether satisfactory if a high degree of carrier stability is required, although for amateur purposes, with careful construction and adjustment, such a simple transmitter offers great possibilities. Crosby has developed a push-pull reactor circuit in which the effect on frequency brought about by changes of supply voltages is arranged to cancel out.

Even so, the push-pull circuit does not compensate for slow frequency drift caused by warming up of the transmitter, valves, etc. The American

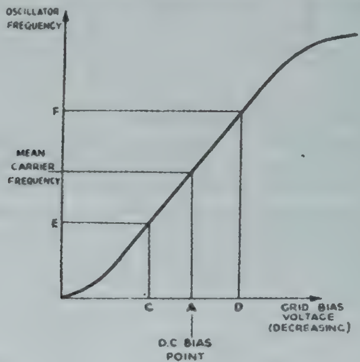


Fig. 82.
A typical reactor characteristic for the circuit of Fig. 81.

General Electric Company has developed a circuit in which automatic control of the mean carrier frequency is obtained by comparison with a stable quartz crystal oscillator. The basic circuit is shown in Fig. 83. The MO—FD—PA part of the transmitter is quite conventional and is such as might be used in a 10-watt 60 Mc/s. amateur transmitter. The reactor is of the type already mentioned, whilst the rest of the circuit provides the automatic frequency control.

Some portion of the output from the PA is fed into a mixer valve where it is combined with a stable reference frequency obtained from a crystal oscillator. The IF from the mixer is selected by a tuned circuit and fed into an amplifier valve. The output from the amplifier is in turn fed into a discriminator or frequency detector, about which more will be said later. It is sufficient here to know that a discriminator is a device which gives an output depending on the frequency of the voltage which is fed to it. The circuit is arranged so that when in tune no output is obtained, but when the frequency of the voltage being fed into the device varies, a positive or negative voltage is

obtained depending on whether the frequency is above or below the " in tune " frequency. A discriminator characteristic is shown in Fig. 90. As the stability of the transmitter is to a large extent decided by the stability of the discriminator a low IF of the order of 500 kc/s. is used where high stability can be achieved. The DC output from the discriminator is smoothed to remove the audio and is then fed back into the reactor grid circuit.

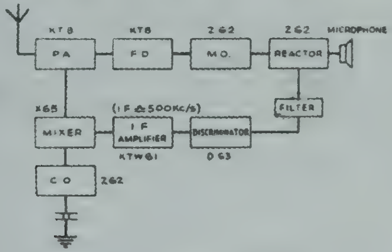


Fig. 83.
A reactance modulator type FM. transmitter with automatic frequency control.

The complete action of the AFC circuit is as follows. Suppose the transmitter frequency drifts higher, then this causes the IF to increase by the same amount and gives rise to a DC output from the discriminator. If this output is of correct polarity to produce a decrease in the oscillator frequency, the transmitter frequency will be corrected until there is no output from the discriminator again. A similar line of reasoning applies to the case of the transmitter frequency drifting to the low side.

In practice a very high degree of stability, approaching that of the best crystal controlled transmitters, can be achieved with such a system.

The Armstrong Phase Modulation System

The Armstrong phase modulator system of producing frequency modulation is designed to achieve a high degree of carrier stability. The system is crystal controlled and produces phase modulation which is turned

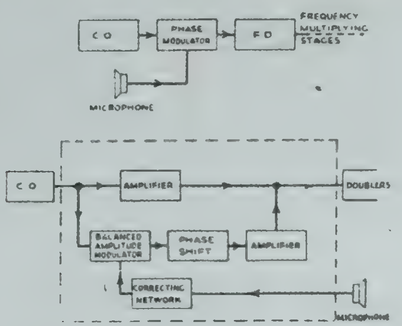


Fig. 84.
The Armstrong system of frequency modulation, above, and details of phase modulator stage, below.

into frequency modulation by means of a simple network in the LF stages of the transmitter. A transmitter using this system has been in operation at Alpine, N.J. (W2XMN), with a power of 40 kW on a carrier frequency of 42·8 Mc/s. The system was flat within 1 db from 30 to 15,000 cycles per second, and the deviation used was ± 75 kc/s.

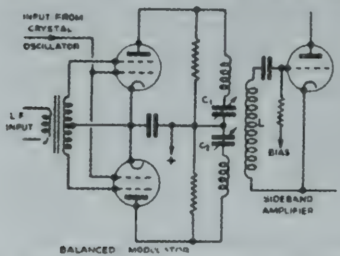


Fig. 85.
Armstrong's balanced modulator circuit.

The elements of the system are shown in Fig. 84. The output from a crystal oscillator is fed into a linear amplifier and also into a balanced amplitude modulator. The latter is a device from which the side-bands only are taken, the carrier being balanced out. The side-bands are then amplified and recombined with the carrier which has undergone simple amplification. The phase of the side-bands is arranged so that when they are in phase with each other they are 90° out of phase with the carrier. The balanced modulator circuit as used by Armstrong in his original experiments is shown in Fig. 85. The condensers C1 and C2 are adjusted to cancel out the carrier frequency. The secondary winding L has a high resonant frequency compared with the crystal oscillator frequency and this provides the requisite phase relationship already mentioned. A vector diagram of the voltages concerned is shown in Fig. 86. The carrier is represented by OA and the side-bands by AC and AD. In the position where the side-bands are in phase with each other and are 90° out of phase with the carrier, it can be seen that a phase shift Θ is obtained.

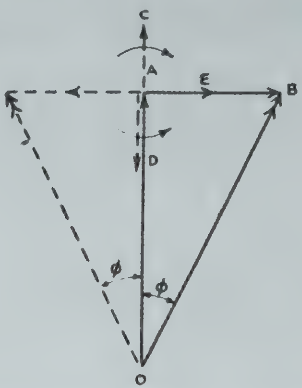


Fig. 86.
Vector diagram of the voltages in the Armstrong phase modulator.

During modulation the side-band vectors rotate about the point A in opposite directions and cause a phase deviation of $\pm \Theta$. It will be seen that some amplitude modulation is also produced, but this is of no consequence as it is removed in the frequency multiplying stages of the transmitter which follow. If the phase deviation is kept below $\pm$ 30° it is found to be substantially linear.

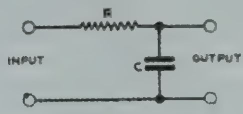


Fig. 87.
Correcting network to produce frequency modulation from a phase modulator.

The equivalent frequency deviation for a given phase deviation is proportional to the modulating frequency, therefore to obtain frequency modulation a network must be inserted in the LF stages of the transmitter, whose output is inversely proportional to the frequency of the modulating voltage. A network such as that shown in Fig. 87 can be used for this purpose. R is a high resistance and C is a capacity whose reactance is small compared with R at the lowest modulating frequency.

The phase deviation at the lowest modulating frequency is limited to $\pm$ 30° for reasons of linearity, and the lowest modulating frequency may be as low as 30 cycles in a high-fidelity broadcasting system. The equivalent frequency deviation in this case is only $\pm$ 15 cycles. If a final deviation of $\pm$ 75 kc/s. is required it is clear that a frequency multiplication of some 5,000 times is required ! It is this need for a large number of frequency multiplications that is the chief disadvantage of the Armstrong system. In practice a fundamental crystal frequency of about 200 kc/s. is used and the frequency is multiplied upwards. After this first set of multiplication stages, the signal is fed into a mixer (which does not alter the deviation) and is changed down to about 200 kc/s. again. More stages of multiplication follow and the process is repeated until the necessary number of multiplications have been achieved to give the correct deviation at the final carrier frequency.

In the case of an amateur type of transmitter the total number of multiplications can be reduced appreciably. Consider a transmitter designed to operate around 144 Mc/s. The audio frequency band which needs to be transmitted for good speech quality is from 300 to 3,000 cycles per second. Assume, too, that a final deviation of some $\pm$ 15 kc/s. will be used. The equivalent frequency modulation deviation at the lowest audio frequency can be calculated to be $\pm$ 150 cycles, calling for a frequency multiplication of only 100 times to give the desired deviation at the carrier frequency. Thus starting off with a 1·44 Mc/s. crystal a simple 144 Mc/s. transmitter could be constructed, the number of valves required being quite small as with modern types it is possible to multiply by as much as five times in one stage.

A Practical FM Exciter Unit Circuit

Fig. 88 shows the circuit diagram of a practical FM exciter unit suitable for driving an amateur VHF frequency modulated transmitter in the 60 Mc/s. band. The EF50 oscillator stage is of the cathode tap type and oscillates at a frequency of about 7·5 Mc/s. The anode circuit is tuned to a harmonic of the oscillator frequency. A total frequency multiplication of eight times is

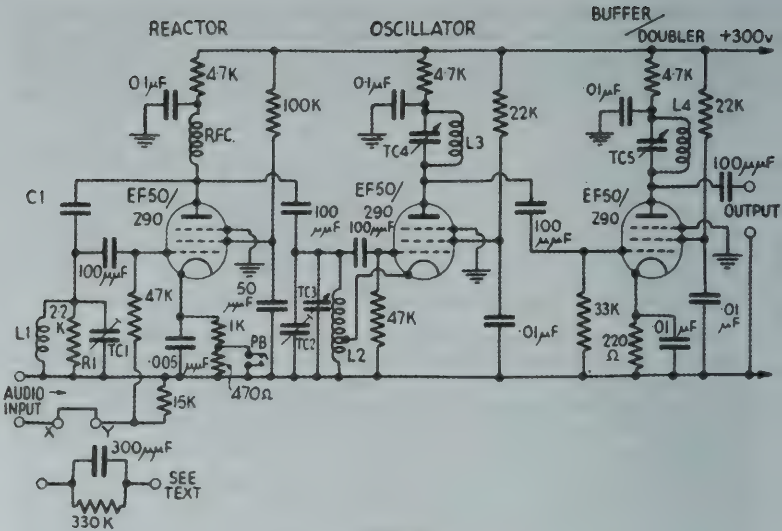


Fig. 88.
A practical frequency modulation exciter circuit.

required to give an output in the 60 Mc/s. band. The multiplier stages are quite conventional and have been described elsewhere. The usual precautions should be taken with the construction of the master oscillator in order to make it as stable as possible. No automatic frequency control equipment has been built into this exciter.

Another EF50 valve is used as a reactance modulator, the 90° phase shift network* being made up of a 2 μμF condenser C1, and the parallel elements L1, R1 and TC1. The audio frequency modulating voltage is applied in series with the 47,000 ohm grid leak. The cathode and screen grid by-pass

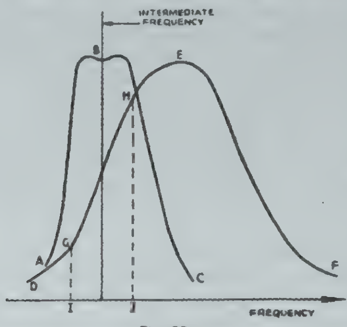


Fig. 89.
An early type of frequency modulation discriminator.

* *G. E. Co., Ltd.* Pat. No. 542,675.

condensers have been made as small as possible in capacity so as to make the dynamic and static performances of the reactor equal. This enables the frequency deviation of the oscillator, due to the reactor, to be measured statically with the knowledge that the dynamic performance using speech modulation will be the same. Methods of frequency deviation measurement are described in a section at the end of this chapter.

The tuning condenser TC1 across the lower arm of the 90° phase shift network may be set up as follows. Insert a milliammeter in series with the oscillator grid leak and press the push button PB, which is arranged so as to short out a portion of the reactor cathode bias resistance. This in turn has the effect of applying a change of potential to the reactor control grid. When the tuned circuit L1, TC1 is correctly tuned there will be no change in oscillator grid current when the button is pressed. On one side of resonance pressing the button will increase and on the other side of resonance decrease the oscillator grid current.

With the circuit arrangements shown, and an oscillator grid current of about 0·15 mA, an audio input of 0·35 volts was required to give a deviation of ± 9·4 kc/s. at 7·5 Mc/s., which after a multiplication of eight times gives

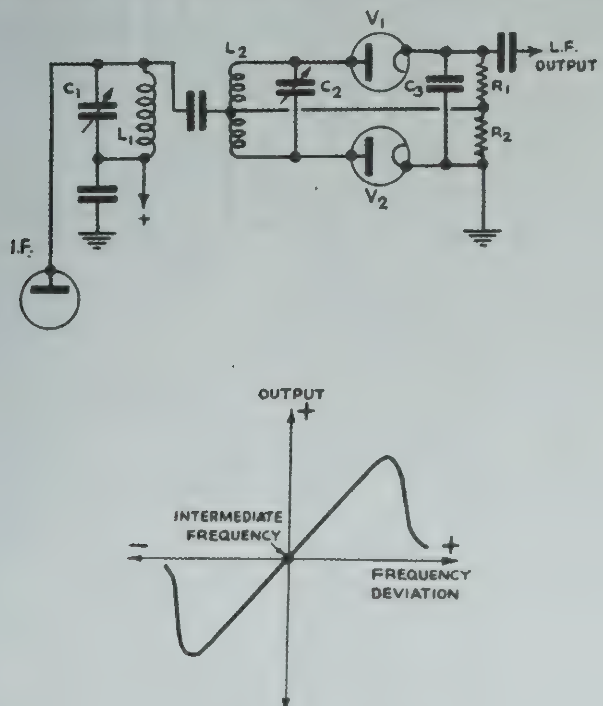


Fig. 90.
The Seeley Discriminator, above, and
a typical discriminator characteristic, below.

a deviation of $\pm$ 75 kc/s. at the output frequency of 60 Mc/s. It should be noted that the deviation which is produced by the reactor is inversely proportional to the oscillator tuning capacity. The circuit shown is suitable for high fidelity work, the response being flat from 50 to 15,000 c/s. For communication purposes the speech band may be restricted to 300 to 3,000 c/s and the frequency deviation may be reduced as desired by reducing the audio input to the reactor as required.

No pre-emphasis is shown in the circuit, but if required the parallel 300 $\mu\mu$F condenser and 330,000 ohm resistance, which gives a pre-emphasis of 100 microseconds, may be inserted between the points X and Y (Fig. 88).

The exciter just described may easily be adapted for 144 Mc/s. operation.

Frequency Modulation Receiver Considerations

An FM receiver differs from an AM receiver in two important ways, first, in the type of detector used, and second in the insertion of an amplitude limiting stage or stages.

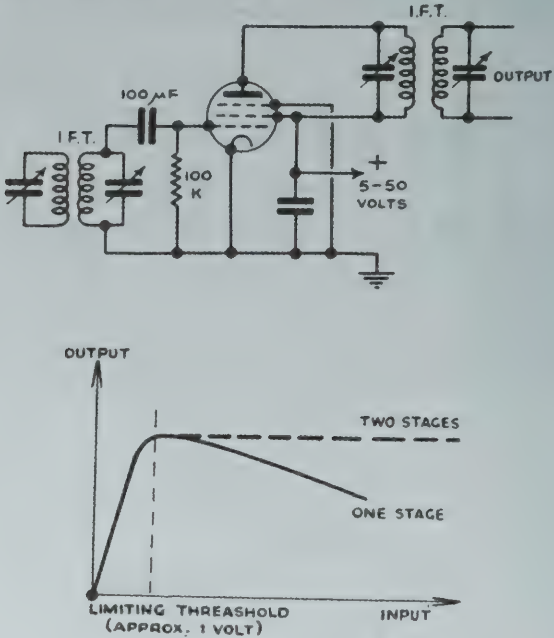


Fig. 91.
A typical limiter circuit, above, and characteristic, below.

Consider the detector, a device which must produce an output proportional to the deviation of the applied signal. The earliest type of frequency detector used a detuned parallel tuned circuit, followed by some form of amplitude detector. Fig. 89 will help to make the method of operation clear. ABC

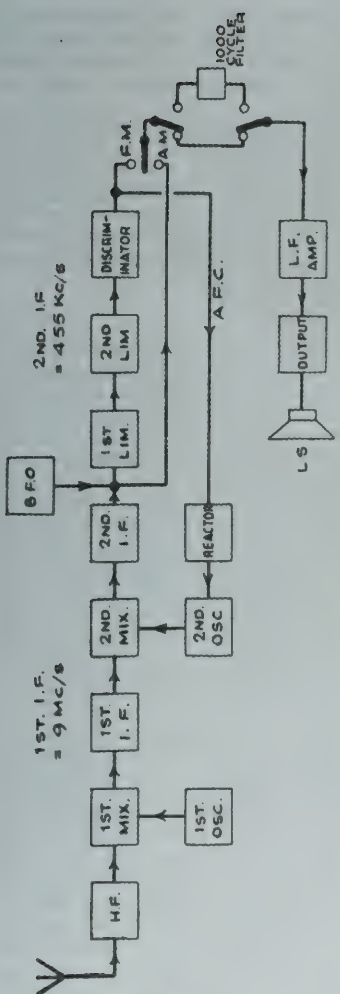


Fig. 92a.
Block diagram of the stages involved in a 144 Mc/s. FM-AM receiver.

VALVES:—			
HF.	Z90	2nd IF.	KTW61
1st MIX.	Z90	Limiters	Z62
1st OSC.	DET20	DISCR.	D63
1st IF.	KTW61	L.F.	DL63
2nd MIX. }	X65	Output	KT61
2nd OSC. }		BFO.	KTZ63
		Reactor	KTZ63

represents the pass band of the receiver IF amplifier, and DEF the response curve of the detuned circuit which follows the amplifier. The tuning of the circuit is arranged so that the straight portion GH is centred on the intermediate frequency of the receiver. Now consider what happens when a signal is applied. As the signal is deviated during modulation the frequency sweeps from I to J—the output from the detuned circuit being proportional to the deviation so long as the deviation is kept within the bounds of the straight part of the curve. The resulting output which is now amplitude modulated can be detected in a conventional AM detector stage. It is this type of detection which takes place when an FM signal is received on an AM receiver by detuning slightly.

The type of frequency detector or discriminator in common use is of the balanced type and was developed by Seeley in America. It has certain noise-reducing properties in itself as when it is exactly in tune it is insensitive to amplitude modulation. The circuit is shown in Fig. 90 together with a typical discriminator characteristic. The circuits L1, C1 and L2, C2 are both tuned to the receiver intermediate frequency and are magnetically coupled as in an ordinary IF transformer. The "hot" end of the primary is connected to the centre tap of the secondary circuit *via* a condenser. The secondary centre tap is also connected to the centre of the load resistances R1 and R2. The ends of the secondary winding are connected to two diode rectifiers. The primary voltage is applied to both diodes, and each half of the secondary voltage is applied to one diode.

The phases and magnitudes of the voltages are such that when exactly in tune the rectified output developed across each half of the load resistance R1, R2 is equal and opposite in sign; thus the sum output across the two diode cathodes is zero. When the signal is deviated, the voltage across one load resistance increases and the other decreases, producing a sum output

across the two diode cathodes which is proportional to the deviation of the signal. The discriminator can be designed to be linear over the range of the deviation of the signal. There are other types of frequency detector circuit but they are in general more difficult to design and adjust and so do not lend themselves to amateur practice.

The discriminator may be tuned by setting an amplitude modulated signal to the middle of the receiver IF passband tuning the primary of the discriminator transformer for maximum output and tuning the secondary for minimum output of the modulating signal from the loudspeaker. Another method is to measure the voltage across the primary of the discriminator transformer, tune the primary for a maximum with the secondary shorted, unshort the secondary, and then tune the secondary until the primary voltage falls to a minimum.

Limiters

The objects of the limiter stage in a FM receiver are (1) to remove amplitude modulation from incoming signals, and (2) to remove noise (which takes the form of pulses of great amplitude). The conventional limiter, which precedes the discriminator, takes the form of one or two stages consisting of pentode valves operating with low screen and anode voltages and arranged with grid leak bias. The screen and anode voltages used are between 5 and 50 volts, depending on the level at which the stage is required to limit, and the time constant of the grid leak and condenser is kept short so that the limiter can handle very quick pulses such as are generated by motor car ignition systems. Time constants of the order of ten microseconds or less are common in practice. The single pentode limiter stage has a characteristic as shown in Fig. 91 which falls off in output as the input is increased past a certain point. This is undesirable and two limiters in cascade are often used, the second limiter being called upon to handle the relatively small amplitude variations which exist in the output of the first stage. In practice, two such limiters give a nearly ideal limiter characteristic as shown in dotted lines in Fig. 91B.

Receiver Design—General

The only other way in which the FM receiver differs from an AM set is in the extra gain which is required to give the best results. The design should be such that the limiters are in operation on the weakest of signals. As many signals used in amateur communication are only just above the receiver noise level it follows that sets should be designed so that the valve and circuit noise originating from the HF stage of the receiver should saturate the limiter valve. As the equivalent noise level at the grid of the HF amplifier of a good receiver is of the order of $\frac{1}{4}$ microvolt and a limiter requires an input of one volt to cause saturation, it is clear that a gain of some 4,000,000 times is required up to the grid of the first limiter ! Using a high carrier frequency not much gain can be obtained in the RF stages of a receiver, leaving all the gain to be obtained in the IF stages. As a high intermediate frequency is necessary on the grounds of image suppression, the achievement of this gain with stability is a difficult problem. It is desirable to go to the double superhet principle for a solution, splitting up the gain between the RF, 1st IF, and 2nd IF stages in order to avoid instability. Such a double superhet receiver is shown in Fig. 92 together with a block diagram of all the stages. The receiver is suitable for the reception of either FM or AM signals in the 144 Mc/s. amateur band. The set will cope with signals using a deviation of $\pm$ 15 kc/s. and the signal-to-noise ratio with an input of one microvolt is 20 db. AM detection takes place in the 1st limiter grid circuit.

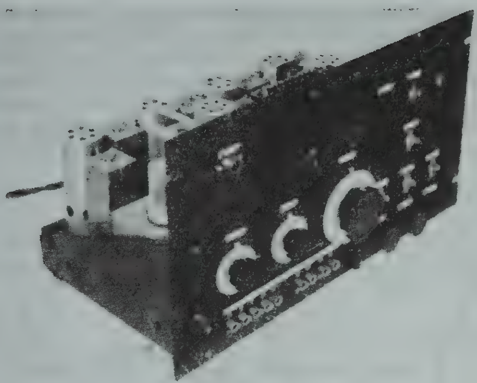


Fig. 92b.
Photograph of the receiver referred to in Fig. 92a.

Noise Reduction in Frequency Modulation Systems

It is the limiter stage in an FM receiver which accounts to a large extent for the reduction of noise in FM systems. All amplitude variations are removed by the limiter. This, however, does not mean that the background of an FM receiver is absolutely silent, for the random-noise components are in effect phase modulated, and so are detected at the discriminator, producing a noise output at the loudspeaker.

Fig. 93 shows the relative noise in an AM and a FM system of the same IF band-width. OC represents half the IF band-width, and in the AM case the noise output would be represented by the rectangle ABCO. In an FM system with a limiter, however, the noise output is represented by the triangle OBC. The reason for this " triangular noise spectrum " is that, the noise being phase modulated, the equivalent deviation is proportional to the audio frequency. If the audio passband is restricted to OF in the diagram it will be seen that the ratio of AM to FM noise will be as area ADFO is to area EFO. It should be noted that the improvement using FM increases as the deviation (and consequently the IF band-width) is increased.

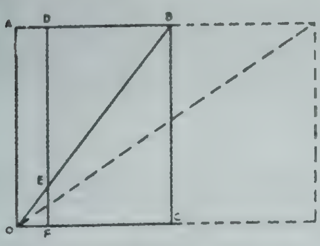


Fig. 93.
The triangular FM. noise spectrum.

In practice there seem to be different optimum deviations for different purposes. For broadcasting where very good signal-to-noise ratios at high carrier levels are required and high audio frequencies have to be transmitted a large deviation of some $\pm$ 75 kc/s. is preferable. Wider deviations could

perhaps be used with advantage but a compromise must be struck with th limited number of channels available even in the VHF region.

For amateur communication purposes where receiver noise usually set the limit of the service area, and many stations have to be accommodated ii narrow wavebands, a narrower deviation will have to be used. A deviatioı of some ± 15 kc/s. seems to be most promising on the grounds of carrie stability, channel width and signal-to-noise ratio.

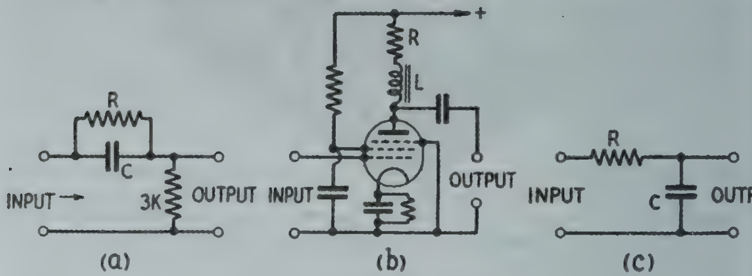


Fig. 94.
Pre-emphasis and de-emphasis circuits.

Pre-emphasis

Pre-emphasis is an artifice which is sometimes used in FM practice in ordeı to improve the signal-to-noise ratio. Pre-emphasis is not peculiar to FM but can also be used on an AM system, but the signal-to-noise ratio improve- ment on FM is greater than can be achieved with AM.

Pre-emphasis is achieved by placing a filter in the transmitter microphone circuit which has the effect of increasing the level of the high modulating frequencies. Another filter which has the reverse effect is inserted in the receiver audio circuit. The filter may take the forms shown in Fig. 94. The value of the pre-emphasis used in a system is referred to as being so many microseconds, the value being numerically equal to the time constant in microseconds. Some common values in use are 50, 75 and 100 micro- seconds.

The reason why pre-emphasis gives an increased signal-to-noise ratio is that it is the high frequency noise components which are prevalent in the output of an FM receiver as can be seen from the triangular noise spectrum of Fig. 93. Thus if the high modulation frequencies are emphasised in the transmitter and de-emphasised in the receiver, then the overall signal-to-noise ratio of the system will be improved. Pre-emphasis is more important in wide-band high fidelity systems than in communication systems where the bandwidth is restricted to 300 to 3,000 c/s.

Measurement of Frequency Deviation

The frequency deviation of an FM transmitter may be measured either statically, applying a variable DC potential to the reactor control grid and noting the carrier frequency shift, or dynamically, using a calibrated discriminator, or by the disappearing carrier method. Particular care should be taken when using the static method of measurement to make sure that the circuit is operating exactly as under the dynamic conditions met in practice. For example degeneration, due to the reactor cathode and screen grid circuits, should not be overlooked.

The measurement of deviation using a calibrated discriminator is very simple. A handy deviation meter may be constructed along the lines of the block diagram shown in Fig. 95. Output from the transmitter under test is mixed with a local oscillator frequency to give an IF of some 10 Mc/s. After amplification the IF signal is fed through one or two limiter stages in order to give a constant amplitude output and then into a stable discriminator. The discriminator should be calibrated statically using a signal generator. An

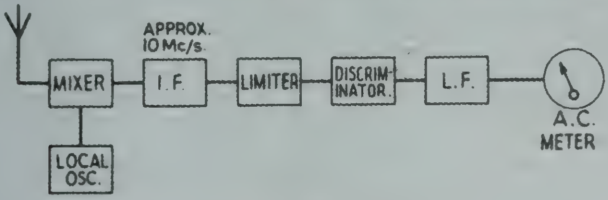


Fig. 95.
Suggested block diagram for a deviation meter.

amplifier of known gain following the discriminator may operate an AC voltmeter whose reading will be proportional to the frequency deviation of the signal under test.

The disappearing carrier method of measuring deviation requires a communication receiver with a very narrow IF filter, any good communications receiver will be suitable, and a variable frequency LF oscillator, such as a BFO, as a source of modulating signal. The method is to tune-in the transmitter carrier, unmodulated, on the receiver. The carrier may be tuned-in to give a maximum reading on the signal strength meter if the set has one, or alternatively the BFO in the set may be switched on and adjusted to give an audible beat note. Next the audio signal feeding the transmitter should be adjusted to the desired frequency. The audio input to the transmitter should then be very carefully increased in amplitude from zero until either the carrier meter reading passes through a minimum, or the audio beat note passes through a minimum, depending on the method of observation used. It will be found that by increasing the amplitude of the modulating signal further carrier disappearances will be found. The order of the disappearance should be carefully noted, starting from zero modulation. The audio signal input corresponding to each carrier disappearance and the frequency of the modulating signal may then be substituted in the formulæ given below to arrive at the transmitter deviation.

1st carrier disappearance.	Deviation	$= 2\cdot4 \times$	modulation	frequency
2nd ,,	,,	,,	$= 5\cdot5 \times$	,,	,,
3rd ,,	,,	,,	$= 8\cdot7 \times$	,,	,,
4th ,,	,,	,,	$= 11\cdot8 \times$	,,	,,

The best modulating frequency to use is something higher than one kilocycle per second. With frequencies below this value it is difficult to separate out the carrier from the sidebands on even the best types of communication receiver.

INDEX

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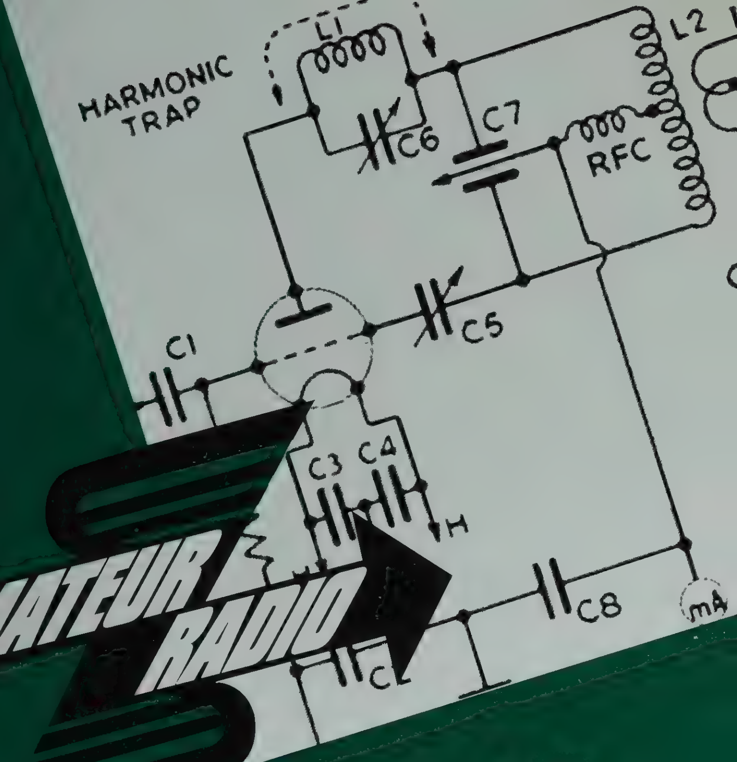

AMATEUR RADIO

TRANSMITTER INTERFERENCE

AN RSGB PUBLICATION

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AMATEUR RADIO

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CONTENTS

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CHAPTER 1 CO-OPERATION AND GOODWILL

WITH the very great increase in the number of radio amateurs now holding transmitting licences, it is inevitable that there should also be an increase in the number of cases of interference to broadcast and television programmes caused by transmissions from amateur stations. It should be realised, however, that the official attitude to the problem is that reception from the two main local broadcast stations, and the local television station, should be free from interference. The receiver must be of reasonably modern design, which excludes the single tuned-circuit valve receiver and the crystal set. Interference to foreign stations and to short-wave reception is regarded by the authorities as a risk which the broadcast listener must accept.

Co-operation Essential

Since it naturally follows that the amateur wishes to live without friction with his neighbours, a great deal of tact, patience and forbearance is necessary in dealing with complaints, especially as some broadcast listeners or television viewers are apt to assume they have a prior right to any reception. It is, therefore, highly desirable that the amateur's attitude should be co-operative and friendly. It is a first duty for him to show the listener that he is anxious to cure the trouble. He should make no attempt to conceal his whereabouts, in fact it is good policy, from time to time, to announce the station address and telephone number over the air with an invitation to anyone experiencing interference to make their difficulties known.

He should be prepared to give interested neighbours an opportunity to see his station in operation, an action that makes for mutual goodwill and one which will serve to dispel the commonly-held, but erroneous, belief that amateur stations work in the broadcast bands. The ability to receive broadcast or television signals free from interference in the amateur's own house should be demonstrated to a complainant. The truth of the old saying that "charity begins at home" is very real in such circumstances. The amateur who has ensured that his own family enjoys interference-free reception, will also have gone a long way towards making this possible for his neighbours. Apart from all other considerations, it is undesirable to bring Amateur Radio into disrepute with the general public.

When a direct complaint is made, prompt attention is most important. The spirit of co-operation will be encouraged if an immediate attempt is made to discover the nature of the trouble, even if a remedy cannot be applied at once. It must be emphasised, however, that, before any investigations are commenced, the amateur should be absolutely certain that his own apparatus is functioning correctly, and is not producing interference due to misadjustment, spurious oscillations, harmonics, over-modulation or key clicks. Only when everything possible has been done at the transmitting end should he consider fitting suppression devices to a receiver.

Even when it can be proved that the remedial measures must be applied at the receiving end—which usually is the case—it is good policy to offer to supply the necessary suppressors free of charge rather than haggle over any questions of responsibility.

The Broadcast Set and the Listener

It is unwise for an inexperienced person to interfere with the internal

wiring of a broadcast receiver or television set. If the amateur is not familiar with this kind of work, he should enlist the aid of an expert. His contacts in the local radio society or R.S.G.B. group should be helpful here.

It cannot be too strongly emphasised that the average listener is not equipped with technical knowledge. A simple, straightforward explanation avoiding technical terms, is essential in dealing with a complaint. . The listener may not understand why it is necessary to fit some device to his receiver, and may be inclined to blame any subsequent fault in his receiver to this cause. This, in turn, may lead to the amateur being asked to effect repairs that have nothing to do with him. For this reason it is important to indicate to the dealer or serviceman who may be subsequently called in to deal with any normal fault, that the receiver has been treated for interference suppression. This will ensure that suppression devices are not removed as " non-standard " additions. It is a good plan to affix a notice inside the receiver stating what modifications have been made and for what reason. This notice should be signed and dated by the amateur concerned.

The Importance of Goodwill

It should be remembered that there are thousands of broadcast listeners to every transmitting amateur. Retention of goodwill is, therefore, of paramount importance. It should never be overlooked that local ill-feeling, which may have little personal effect on the amateur himself, can often make things very unpleasant for his wife or other members of his family.

In the following pages the various forms of interference are dealt with in detail together with the appropriate curative measures. Very few cases of interference are incurable given patience and goodwill.

Finally a word to the newcomer to Amateur Radio. Do not stay off the air because you think you will cause interference; this is a negative and unworthy attitude. From the very start make every effort to ensure that anyone in trouble comes to you immediately and does not wait until infuriated beyond bounds, when he will probably regard you as a menace. In really difficult cases, do not hesitate to call upon the good offices of the Interference Section of the G.P.O. Radio Branch, which exists to help both sides and is always ready with good advice and practical help.

CHAPTER 2 INTERFERENCE TO MEDIUM-WAVE BROADCAST RECEPTION

FOR convenience the main types of interference to receivers working on the normal broadcast frequencies may be classified into five groups. Remedies will, of course, vary considerably with local conditions and no general cure can be claimed as effective in all cases.

Causes of Interference

The five main types of interference are those caused by:

1. Key clicks, thumps and spurious emissions, such as parasitic oscillations, over-modulation and harmonics, all of which are curable only at the

transmitter. Most of these, incidentally, also come under the heading of "Transmissions on other than the authorised frequency band."

2. Blanketing, which may mean anything from a rather tiresome variation in the volume of the broadcast programme, to a bad case where the programme entirely disappears.

3. Cross-modulation which results in the appearance of an unwanted signal only at points on the tuning scale where strong broadcast programmes also appear; the interference not being audible over the remainder of the tuning range.

4. Break-through, which exhibits itself as the appearance of an unwanted signal over most, if not all, of the broadcast band. Although interference is severe it does not usually affect the volume of the broadcast programme.

5. Second channel and superheterodyne interference, which shows up as tuneable appearances of an unwanted signal at certain points on the broadcast band, not necessarily associated with any other station.

It will be appreciated that more than one type of interference may be present at any one time and, further, that these may reach the receiver by one or more routes, or a combination of several routes. In general, however, interference arrives at the receiver either by way of the aerial or through the mains connection. Sometimes it may be picked up by the actual set wiring but this is relatively unusual.

If interference is entering the receiver by way of the aerial or earth leads only, then it will cease as soon as these leads are removed from their sockets. Such interference may be excluded by the interposition of a suitable filter. (See Chapter 4.)

A Typical Case

In the meantime let us assume a typical case. A complaint is received—"Do you know you are blotting out our wireless?" It is often couched in more forceful terms, but the essentials are the same every time.

After checking the transmitter, a visit should be made without delay to the complainant's home and arrangements made for the transmitter to radiate during the visit. Ways and means of doing this satisfactorily are discussed in a later chapter.

Particularly where interference is not due to telephony and, therefore, not clearly identifiable, it should be borne in mind that it may not be due to the amateur station at all. If this is so, a clear and reasonable explanation should be given. If the interference is definitely identifiable and is due to the amateur station then it should be admitted at once. Any attempt at bluff or denial only makes things more difficult for everyone concerned, especially when the aggrieved listener calls in the Post Office. It is also worth remembering that the average listener will not appreciate derogatory remarks regarding either the age, selectivity or appearance of his receiver. You may think it is an antique relic. It is a mistake to say so!

Tracing the Cause of Interference

Having identified the interference, it should then be classified, if possible, into one of the five groups mentioned earlier, and the point of entry ascertained. First remove the aerial and then the earth leads and note the result. If the interference ceases the answer is obvious. If it does not, or even increases, it is probably arriving *via* the mains lead. An increase of volume being due to A.V.C. action caused by the removal of the aerial lead.

As mentioned previously, more than one type of interference may be present and more than one mode of entry involved, but the various forms

are easily distinguished from each other, and it is the predominant one which should be tackled first.

Tests such as those mentioned above are a matter of routine. Curing the trouble is a matter of careful, systematic and progressive elimination.

Mains borne interference may be due to feed-back through the mains wiring at the transmitter itself or, more often, to unscreened mains wiring in the vicinity, acting as an aerial, picking up energy radiated by the transmitting aerial. This is very prevalent in many modern housing estates where much of the house wiring is carried out with unscreened cab-tyre cable.

It is at this point that the difficulty of dealing adequately with all the possible combinations, in a publication of this type, becomes apparent. In Chapter 4 will be found a detailed selection of typical cases of interference and the measures taken to cure the trouble. Intelligent application of these cures should see most amateurs safely through their difficulties so far as broadcast interference is concerned.

CHAPTER 3 INTERFERENCE CURABLE AT THE TRANSMITTER

KEYING is the chief cause of interference from a telegraphy transmitter, but this trouble may usually be cured by one or more of the devices to be described. Interference from this source takes the form of clicks or thumps, sometimes accompanied by hum, in the neighbouring receiver.

Interference from telephony is not curable at the transmitter, unless it is due to misadjustment or over-modulation. In this connection it is desirable to point out that the modulation level should always be kept low during the main broadcasting hours in order to ensure that the peaks of speech do not over-modulate. Again emphasis is placed on the need for the amateur to ensure that his apparatus is correctly adjusted and that no over-modulation is taking place.

Key-Thump

Key-thump may be cured in a number of ways depending on the method of keying used. One of the most common places to insert the key is in the cathode circuit of one of the early stages of the transmitter, and occasionally in the oscillator circuit itself. This later method produces a more severe click at " break " than at " make " and is strangely difficult to eliminate completely. It may, however, be reduced considerably by employing the

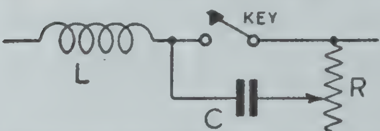


Fig. 1.
A thump filter for keying in D.C. leads. The inductance L is usually iron-cored.

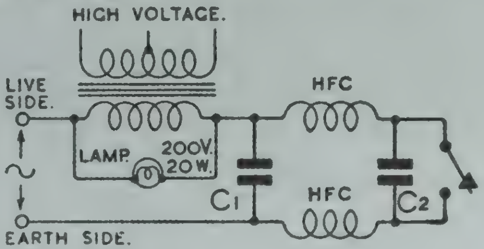
arrangement shown in Fig. 1, which consists of an inductance L in series with the key, and a condenser C and resistance R in parallel with it. The inductance will reduce the clicks on " make " and the condenser-resistance

combination at " break." The inductance should be any value up to 50 Henrys with no current flowing, and the value of C varied between ·1 and ·5 μF. R can be between 25 and 500 ohms according to C. This thump-filter can also be used with success if anode circuit keying is employed.

Primary Keying

The practice of keying in the primary circuit of the H.T. transformer has been found to be very successful in eliminating clicks, but is apt to give " tails " to the Morse characters unless the smoothing condensers are kept to small values. A fairly heavy " bleeder " current will also help. Some load should be connected across the primary to stop heavy surges. A primary keying arrangement is shown in Fig. 2.

Fig. 2.
Primary keying.
C1 0·01 μF to 0·05 μF.
C2 0·001 μF.
HFC 210 turns No. 30 D.S.C. on 5" former.



Grid Blocking Keying

The method known as grid block keying is illustrated in Fig. 3. When the key is " up " a large negative bias is applied to the grid of the valve being keyed thus shutting off all anode current. In the " down " position this additional bias is shorted-out through a high resistance ; a value above 100,000 ohms will be satisfactory.

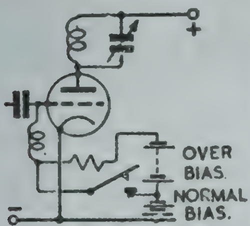


Fig. 3.
Circuit of Grid blocking keying.

Valve Keying

An excellent clickless keying circuit is shown in Fig. 4, where V2 is used as the keying valve. When the key is " down " this valve will pass the cathode current of V1 with only a few volts drop. Although the circuit shows directly-heated valves, it is equally suitable for use with indirectly-heated types. If a directly-heated valve is used for V2 then a separate filament supply is necessary. Several methods of valve keying are possible, but the arrangement described has been found to be free from click.

Use of Relays

It should be borne in mind that when using a relay for keying, a thump or click will take place on the make or break of the relay contacts, and that a further click will occur as the key itself completes the circuit through the energising coil of the relay. The filter shown in Fig. 1 should overcome this trouble, but it may be helpful also to include two R.F. chokes in series with the leads to the relay contacts, close to the relay. Fig. 5 shows this arrangement.

Trial and Error

Since no exact values can be given for any of the suppression devices described, owing to each case requiring individual treatment, a system of trial and error must be adopted and this often calls for considerable patience.

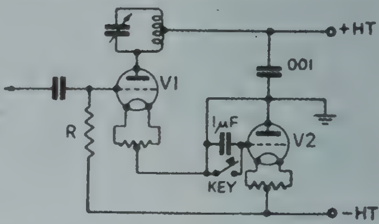


Fig. 4.
Valve keying circuit.
V1 may be either an oscillator or buffer stage.
V2 may be any low impedance triode capable of passing the current required by V1.

Interference from parasitic oscillation can take place but is usually not serious. Any of the normal forms of anti-parasitic device can be used, including R.F. chokes, shunted by a low resistance in the anode circuit,

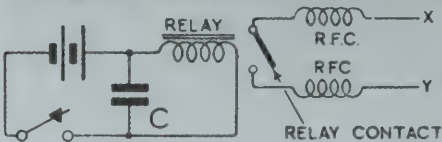


Fig. 5.
Interference suppressors in a relay keying circuit.

close to the anode of the valve. The chokes may conveniently take the form of a few turns of 18 S.W.G. wound round, but not touching, a 50 ohm resistor of suitable wattage. Care should always be taken when neutralised circuits are used to ensure that oscillation does not occur at any time.

CHAPTER 4 INTERFERENCE CURABLE AT THE RECEIVER

THIS Chapter deals in detail with the types of interference classified as Groups 2, 3 and 4 in Chapter 2. It is again emphasised that several types of interference may be present simultaneously. For this reason do not be discouraged if success is not achieved immediately.

Blanketing

Interference due to blanketing usually affects the volume of the received programme and is almost always due to the unwanted signal appearing at the grid of an early valve in the receiver, by way either of the aerial or earth lead.

It may be that either the selectivity of the receiver is poor, or that, although reasonably good for broadcast reception, it is not good enough to deal with a very high signal-input from a transmitter only a few yards away. Removal of the aerial or earth lead usually stops the interference or reduces it to a

very low level, and a permanent cure may be effected by the insertion of a series filter in the aerial lead.

The simplest cure in cases where interference is not severe, is to insert a small series condenser in the aerial lead close to the receiver. This has the effect of reducing the coupling of the aerial to the set, and provided the gain of the receiver is sufficient no appreciable difference in broadcast reception will be noticed. The capacity should be between 250 and 500 $\mu\mu$F, the best value being found by experiment.

In cases where a long outdoor aerial is in use, it is often advantageous to shorten it considerably. The insertion of a small series condenser will do this electrically if the owner does not wish the aerial to be cut.

In this connection, it is wise to give some thought to the transmitting aerial. It should be placed as far as possible from any broadcast receiving aerial, and if possible at right angles to it. Needless to say if feeders are used they should be carefully matched to avoid radiation.

For medium cases where wipe-out is not complete, an untuned series filter as shown in Fig. 6A, will usually clear the trouble. A convenient filter may be made by winding a piece of wire a ¼ wavelength long, on a former ½″ to ¾″ in diameter, and connecting this in series with the aerial of the receiver as near to the aerial plug as possible.

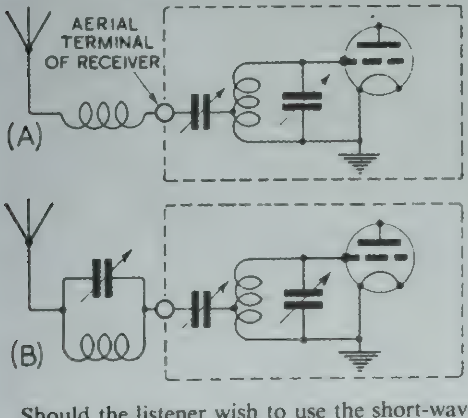


Fig. 6.

Showing how an untuned (A) or tuned (B) wave-trap should be connected between the aerial and the first tuned circuit of a broadcast receiver.

Should the listener wish to use the short-wave bands—very few ever do—or should the interference be really serious, a tuned filter of the type shown in Fig. 6B should be used, and arrangements made to switch it out of circuit when the short-wave range is in use. Such filters are most effective when the coil is large and the value of the condenser relatively small. Treatment of this type of interference is the same whatever the frequency of the transmitter.

For more serious cases still a low-pass filter may be inserted in the aerial lead, again close to the receiver. The circuit is shown in Fig. 7. This filter is more suitable for removing interference caused by 1·7 Mc/s. or 3·5 Mc/s. transmissions, than for the higher frequency bands.

It is essential that the coils be wound with No. 30 D.C.C. wire with the turns touching and secured in place with Durofix or similar cement so that they do not become loose. They may be wound on valve bases, or any other 1⅜″ diameter former. The condensers must be of a reliable make and of the mica type. Winding details for the coils and the capacities of condensers

required for various bands are given in Table I. It should be noted that the filter for the 1·7 Mc/s. band can be used on the 3·5 Mc/s. band, but that the 3·5 Mc/s. filter is not suitable for use on 1·7 Mc/s. transmissions.

TABLE I.

Coil	Turns 1·7 Mc/s.	Turns 3·5 Mc/s.	Turns 7·0 Mc/s.
L1	20	20	13
L2	30	30	20
L3	27	13	9
C	·0002 μF	·0002 μF	·0001 μF

The filter is arranged to have a characteristic impedance of 400 ohms at each end, as this figure is likely to suit most aerials and receiver input circuits. The loss at broadcast frequencies is negligible.

If the interference disappears only when the earth lead is disconnected, it is probably a case of coupling due to a common earth connection, *i.e.* both transmitter and receiver connected to the water pipe. Use of a separate earth for one or the other will effect a cure.

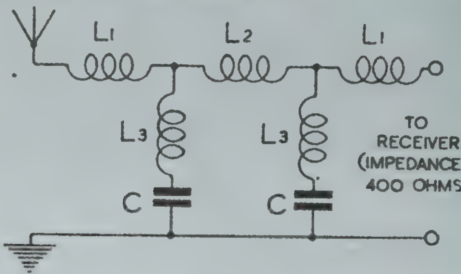


Fig. 7.
A low-pass filter which can be added to a broadcast receiver to eliminate interference.

Cross Modulation

The primary cause of this type of interference is similar to the previous form of "blanketing," but only occurs with a modulated transmitter. It can easily be recognised by the fact that the speech appears on top of all or most of the broadcast stations when tuned in, but does not appear between stations.

It is caused by a severe overload of the first R.F. valve of the receiver, which, due to this, commences to rectify signals from the local transmitter. When the receiver is tuned to a station the broadcast carrier is modulated by the rectified signal already present in the R.F. stage and as a result is passed to the detector stage together with the broadcast signal, and consequently both appear in the output. It will be seen, therefore, that the presence of both signals simultaneously indicates this form of interference.

The cure is to reduce the amplitude of the signal from the amateur transmitter in the first R.F. stage of the broadcast receiver, and any of the remedies

previously suggested may be used. In addition, since this trouble more often occurs in a receiver having an R.F. valve with non-variable-*mu* characteristics the substitution of a variable-*mu* valve in this stage may improve matters. A further cure can often be effected by the connection of an R.F. choke direct to the grid of the first R.F. stage in series with the lead thereto. It may be necessary to screen this lead.

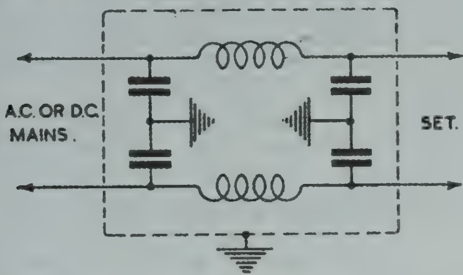


Fig. 8.

Interference suppressors in the mains leads to transmitter or receiver. The earthed metal box containing the filter is a precaution against direct pick-up or radiation by the chokes.

Break-Through

The audible effect of this type of interference is the same as for blanketing, but it is unusual for the interference to be so strong or for it to affect the strength of the signals received from the broadcast station. This is because the point of entry is usually by way of the mains lead and very occasionally by way of the wiring of the receiver itself. In the first place, therefore, a capacity of between ·0001 and ·001 μF should be connected from the " live " side of the mains socket to chassis. If this cures the trouble, it is unnecessary to go further, although an equally effective cure can usually be made by connecting the condenser from one side of the heater line to chassis. The condenser may then be an ordinary low-voltage paper type as it is connected across only a few volts A.C. If it is necessary to connect the condenser across the mains, it should be of good quality with mica insulation. In the case of a receiver suffering from 28 Mc/s. interference this lead to chassis must be as short as possible. Connection of the condenser by means of a lead several inches long may be useless, whereas a complete cure may often be effected by cutting this down to half an inch.

Many broadcast receivers have the various decoupling condensers and other " earthy " leads joined to an earth wire which may be anything up to a foot long connected to chassis at one end only. At frequencies above 28 Mc/s. in particular, an amateur soon realises that the opposite end of a piece of wire earthed at the other end is far from being at earth potential to R.F. ! By the simple expedient of connecting such a lead to chassis at more than one point, serious interference can often be eliminated.

Should interference still persist from this source, a more elaborate filter should be fitted in the mains lead. A suitable circuit is shown in Fig. 8. Good quality condensers only should be used, and care taken with the insulation. The filter must be screened in a metal box which should be earthed, and connected as close as possible to the receiver, otherwise the lead between the filter and the receiver may require screening.

Cases have been known where interference of this type still persists even after a mains filter has been fitted.

This points to pick-up by the set wiring, and may occur at the most unlikely places. On A.C. or battery sets the offending stage may be found by the simple process of removing the valves in turn beginning at the R.F. amplifier (if any), to the power valve, noting when the trouble ceases. Rectification can take place even on the grid of the output valve, a 5,000 ohms carbon resistor in series with the grid and a ·0001 μF condenser from grid to cathode usually effects a cure. Incidentally this is an excellent cure for R.F. in modulators and audio amplifiers of all types, except Class " B " output stages.

In one receiver R.F. was being fed into the circuit due to a rather long gramophone pick-up lead. A ·0001 μF condenser connected to chassis overcame the trouble. The insertion of an R.F. choke in the grid lead to the detector stage, close to the valve, may also assist to cure this form of interference.

Second Channel and Super-Heterodyne Interference

It is perhaps not correct to include this trouble under the heading of interference curable only at the receiver. It is, however, a receiver fault, and not due to radiation by the amateur station on an unauthorised frequency.

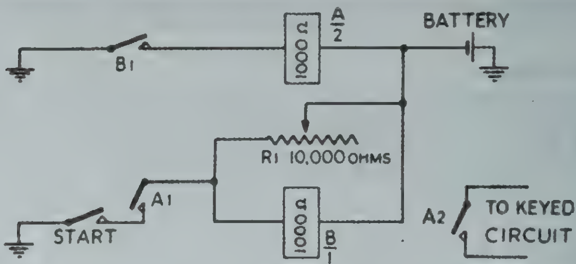


Fig. 9. Circuit of Auto-Keyer.

Due to the choice of intermediate frequency in the receiver it is possible for the amateur station to produce a signal at the I.F. of the receiver by beating with the oscillator frequency or alternatively by some other combination of strong signal or harmonic either of the amateur signal or super-het. oscillator. It may be necessary to screen leads in the receiver or to take other steps to prevent the signal from the transmitter combining with the local oscillator of the receiver.

Probably the easiest course is to find, and then avoid, the spot frequencies on the appropriate amateur band which cause the trouble.

Remedial measures are often a matter of progressive reduction. Attention to the points already mentioned will usually be all that is necessary, but in a difficult case, anything which reduces interference and does not materially affect the broadcast reception should be incorporated until interference disappears. Success has, before now, attended the completely unscientific course of tapping around with a chassis-connected ·0001 μF condenser to find points in the wiring of a receiver where interference can be reduced.

In general, however, there is nothing in the problem of broadcast interference once it is faced and tackled fearlessly. No two types of broadcast receiver are alike and some seem to be designed in order to render servicing

as difficult as possible; some even hardly appear to be " designed " at all ! Good clean layout, accessibility and short earth returns are most important. If all manufacturers would by-pass the heater line on their receivers in the manner suggested, much of the trouble with broadcast interference from amateur transmissions would never arise.

Automatic Key to Send Dashes

Should it be necessary to arrange for the transmitter to be keyed during broadcast interference investigation, the simple unit shown in Fig. 9 will prove very useful. The device will cause the transmitter to send dashes of varying length, according to the prearranged position of R1 which acts as a release delay slug on relay B.

The operation is as follows : When the " start " switch is closed, relay B is energised and contact B1 energises relay A. A2 operates the keying circuit and A1 opens, de-energising relay B. Relay B, however, releases slowly due to the slugging effect of R1. When B releases, B1 releases relay A. Contact A1 restores, re-energises relay B and so the cycle continues.

Telephony may be simulated by receiver feed-back to the microphone or by passing the output of an audio oscillator through the modulator. The transmission of modulated or unmodulated signals may be controlled by pre-arrangement with someone left in charge at the transmitting end so as to reduce transmission to an absolute minimum. The station call-sign should be announced at the end of each test transmission and the reason for the test made clear. This announcement should, of course, be made by the licencee. If the complainant and the amateur are both on the telephone, tests can be made more easily and transmissions kept to a minimum.

Relays suitable for the test keyer are available from ex-Government surplus and will work quite satisfactorily from a 9 volt grid-bias battery for several hours.

CHAPTER 5 INTERFERENCE TO TELEVISION

INTERFERENCE to television reception is without doubt the most important problem at present facing the radio amateur and the radio industry. Due to its special nature, the problem requires special treatment, although most of the general remarks which concern broadcast sound interference apply, especially those which emphasise the need for co-operation and immediate attention to a complaint.

Owing to the rapid growth and extending hours of television, more and more radio amateurs are becoming aware that they will have to take steps to deal with the trouble if they are to retain the freedom of operating times they at present possess.

Interference generally takes the form of distortion of the picture, or a series of diagonal lines across it. Sometimes it is a trellis pattern, but in severe cases a complete blanketing occurs leaving the screen completely white.

Typical examples of pictures with interference such as could be caused by an amateur transmitter are illustrated in Figs. 10 and 11. From these it will be seen that even a small interfering signal can cause a very noticeable deterioration in the picture, and it can be assumed that any interference

above about – 40 db compared with signal level, is likely to cause trouble, and steps should be taken to reduce or remove it.

Harmonic Radiation

Generally, more interference troubles occur on the outskirts of the service area, due to the lower field strength of the vision signal. The most probable source of interference is, of course, harmonic radiation, from either the P.A. or driver stages of a neighbouring amateur transmitter, since certain of the amateur bands show harmonics in the Television range of frequencies. Table II, which illustrates this point, also indicates the bands where trouble may occur even when it would seem that no radiation in the Television band is taking place.

Fig. 10A.

In the case of transmissions in the 28 and 58 Mc/s. amateur bands it is possible that interference may be caused by harmonics from a 14 Mc/s. stage in earlier sections of the transmitter. This applies particularly to the 58 Mc/s. band since such harmonics would be nearer the 45 Mc/s. Television band than if the transmitter were working on a fundamental frequency in the 14 Mc/s. band. These harmonics can pass through the P.A. stage without serious loss, and can be radiated from the aerial.

Fortunately the radiation of harmonics is fairly easily controllable at the transmitter. With this in mind, amateurs should take steps to ensure that harmonic suppression devices are incorporated into any new apparatus they may build. Details of suppression methods are given later.

Television Receiver Design

The second, and more serious, cause of interference is due to certain features in television receiver design; such troubles cannot be cured at the transmitter end. The effect of this form of interference is that, although the amateur is transmitting on his licenced frequency, and is not radiating harmonics in the Television band, his signals appear in the sound or vision channels, or both, of the nearby receiver.

Some receivers have a local oscillator operating on 37 Mc/s. and an I.F. amplifier centred on

Fig. 10B.

Two television pictures showing the effect of C.W. interference at a level of (A) 0 db and (B) –10 db compared with signal level. The general effect being the black diagonal lines across the picture.

8 Mc/s. The second channel image frequency is, therefore, 29 Mc/s.—the centre of the 28–30 Mc/s. amateur band. The selectivity of the R.F. circuits at this frequency is totally inadequate to deal with the high field-strength of a local amateur transmitter.

The most satisfactory solution in this case is the insertion of a high-pass filter in the receiver aerial lead so as to attenuate severely signals between 28–30 Mc/s., while offering no deterrent to signals in the Television band. Suitable designs for such filters are given later.

Again, some receivers have I.F. amplifiers covering the 7·0–7·3 Mc/s. amateur band, and it requires only a very small amount of unscreened wiring between the mixer anode and the first I.F. grid to pick-up R.F. from a nearby amateur station working in that band. Certain makes have I.F.'s in the 14 Mc/s. band, and similar trouble can be caused by stations working in that band. Signals on these frequencies can be picked up by the aerial and passed through the mixer stage, reaching the I.F. stages in that manner. It is possible also that the feeder may accidentally be of such a length as to resonate with the 7 Mc/s. or 14 Mc/s. signals.

Fig. 11A.

Where interference is due to such causes, it would seem that, in the event of filters not proving satisfactory, screening the leads mentioned above is necessary, but this cure should not be attempted by anyone not really familiar with Television receivers. A mains filter and a high-pass filter in the aerial will also assist in overcoming this type of trouble.

Harmonics from Communication-Type Receivers

One further source of interference that occasionally occurs is due to harmonics from the oscillator stage of an amateur's own communication receiver. Such harmonics are sometimes sufficient to cause considerable trouble. Mains filters and further screening of the offending section of the receiver are likely to provide a cure.

Television Aerials

One point that should not be overlooked when investigating cases of interference is the efficiency of the Television aerial. In areas fairly close to the vision transmitter it is not unusual to find that the owner

Fig. 11B.

Illustrating two examples of interference having reached the synchronising circuits; (A) with a level of –20 db, and (B) with a level of –30 db compared with signal level.

of a perfectly good vision receiver is using for an aerial an odd piece of wire, or at best a quarter wavelength of wire, fitted up indoors. Such an arrangement is much more prone to interference than a properly designed outdoor aerial with a correctly matched feeder system. The latter arrangement should always be recommended in cases of interference. If it is argued that such an aerial gives too great a signal, the remedy is to introduce a simple attenuator between feeder and receiver. Devices of this nature are frequently supplied by a firm when erecting an aerial. Details of attenuating pads having various degrees of attenuation are given in Fig. 12. The correct amount of attenuation should be found by trial and error.

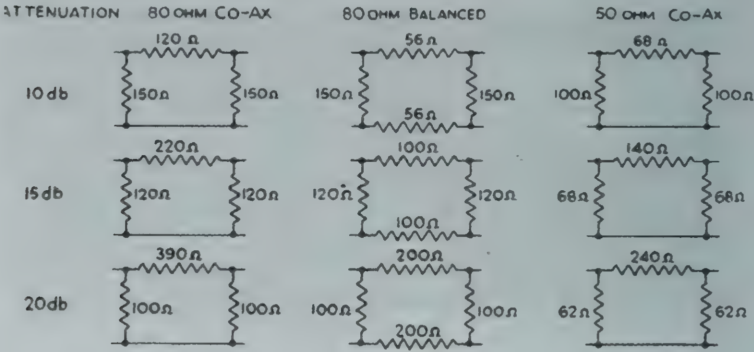


Fig. 12.

Details of attenuators for use in vision receiving aerials. For use see text. 80 ohm co-axial or balanced types are usually suitable for 70–80 ohm cables used on a single dipole. 50 ohm types are for use with the 50 ohm cable used with the dipole and reflector type of aerial. All values are taken as nearest 5 per cent. values on standard logarithmic numbers.

A good Television aerial without an attenuator may work too well and actually introduce interference, at the same time overloading the receiver with the Television signal.

Band. Mc/s.	Harmonic.	Range. Mc/s.	Chance of Interference.
1·715–2·0	21st, 22nd, 23rd, 24th	36–48	Negligible
3·5– 3·8	12th, 13th	42–49·4	Slight
7·0– 7·3	6th	42–43·8	Moderate
14·0–14·4	3rd	42–43·2	Serious
21·0–21·4	2nd	42–42·8	Serious
28·0–30·0	2nd	56–60	Serious
58·5–60·0	—	Fundamental	Serious

TABLE II.

Shows the position for the frequencies used by the London Television Station.

Attenuators

Attenuators should be constructed in a metal box and suitably earthed, although this will be automatic if concentric sockets are used for the input and output connections to the aerial feeder and set input. In some cases it may be found that the fitting of an attenuator will cure interference troubles, since naturally the interference will be attenuated as well as the signal.

Suppression Devices

Before dealing in detail with the practice of curing interference it is desirable to consider the various points that are likely to require attention in a Transmitter and Television receiver. Fig. 13 shows a schematic arrangement of a transmitter, and indicates these points, while Fig. 14 shows the receiver.

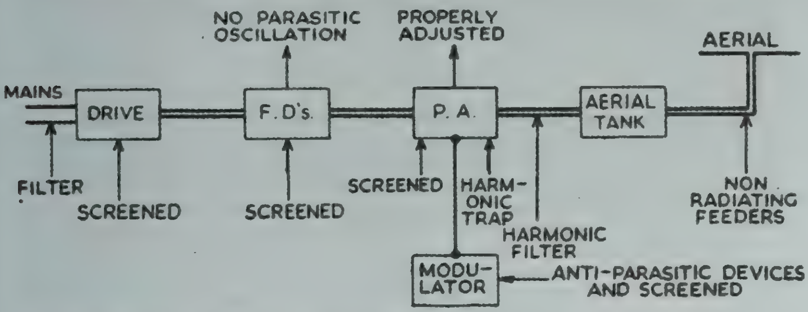


Fig. 13.
Schematic diagram of Transmitter showing points requiring attention to prevent radiation of harmonics.

Commencing with the transmitter, as the source of interference, a mains filter should always be installed. Due attention should then be paid to the screening in all its stages, followed by the incorporation of a harmonic rejector in any of the anode circuits which may require it. If the screening is properly carried out this should be necessary only in the P.A. stage. The permanent installation of a harmonic indicator in this part of the circuit is recommended, and details of such a device are given on page 22. In this

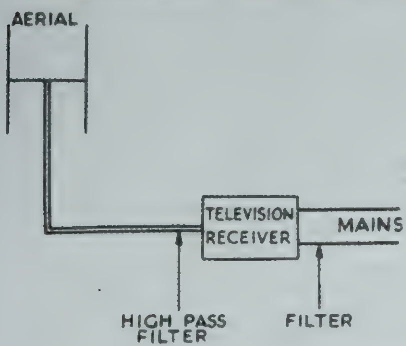


Fig. 14.
Schematic diagram of Television receiver showing points requiring attention to prevent pick-up of unwanted signals.

manner it will always be possible to keep a check on the harmonic radiation, and adjustments made as necessary when the frequency is changed. Finally, consideration should be given to the use of a link-coupled aerial circuit with a filter incorporated in the link line.

It is not always necessary to employ both a harmonic rejector and a filter in the aerial circuit, but in cases of severe interference and a low field strength from the vision transmitter it may be found desirable.

In certain instances the modulator, if of the Class B type, may require anti-parasitic devices.

At the receiving end a high-pass filter can be installed in the feeder line from the television aerial—provided it is correctly designed—and mounted close up to the receiver, whilst a mains filter should remove stray and slight interference that may be entering the receiver that way. In addition, the direct pick-up type of interference mentioned previously may necessitate modifications to the receiver. Attenuator pads as described may also be fitted.

Drive Frequency

Where transmission is required in the 28 Mc/s. or 58 Mc/s. band it is advisable to arrange that the frequency of the driving system, whether V.F.O. or crystal, shall be in the region of 5 Mc/s. or 10 Mc/s., so that the required frequency in the P.A. can be obtained by the use of a tripling stage for 10 to 30 Mc/s., thus avoiding the generation of any frequencies near the 45 Mc/s. band. This has proved completely efficacious in some cases.

The popular tri-tet oscillator is a prolific generator of harmonics and is extremely difficult to suppress, because much radiation can occur from the cathode coil. For this reason its use is not recommended where Television interference is likely to occur.

Suppression of Harmonics

It cannot be too strongly emphasised that it is of little use attempting to suppress the radiation of harmonics from a transmitter if it is poorly built, *i.e.* without proper earthing and screening, or is improperly adjusted. Parasitic oscillation, inaccurate neutralisation, as well as over-modulation, are all symptoms of bad adjustment. As the subject of transmitter construction and adjustment is dealt with in other Society publications, it is not proposed to consider it here.

If interference troubles are being experienced it may be well worth while to reconstruct the transmitter or at any rate the P.A. section of it, bearing in mind the points mentioned, especially the screening.

It should be remembered that the harmonics from a V.F.O. or crystal oscillator are often sufficiently strong in themselves to cause interference, and although the stage itself may be screened it must be recognised that the power supply leads to it require filtering and possibly screening. This applies especially when the unit is supplied from a power pack remotely situated, as is often the case with a V.F.O. where it is customary to have it close to the receiver and connected to the transmitter or to its own power pack by long leads. The leads from the power pack could quite easily be of such a length as to resonate with harmonics falling in the Television band, thus acting as an aerial.

A loosely coupled aerial circuit should always be used, preferably the type using a form of link coupling to the aerial tank circuit, which should be separate. A short length of concentric line is very convenient for this purpose, since it permits the easy installation of a filter, as described on page 19.

Fig. 15 illustrates an arrangement of the P.A. tuned circuit and a link-coupled aerial circuit. The advantage of this method is that any type of aerial can be coupled to the aerial coil, without any further adjustment or alteration being made to the harmonic filter, since it remains in an unaltered part of the aerial coupling system.

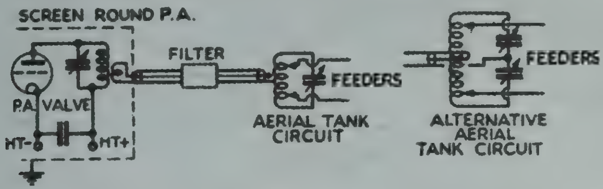


Fig. 15.
Shows a typical link coupled aerial circuit and how a filter can be installed. The link line may be any length within reason.

In the case of an aerial using low impedance twin (80 ohm) feeders, the separate tank may be dispensed with and the filter installed directly in the feeder line. Generally, however, it is a definite advantage to use the loose coupled system. Fig. 16A shows the arrangement of a filter suitable for a low impedance balanced line. Fig. 16B shows a re-arrangement with both networks in the same line.

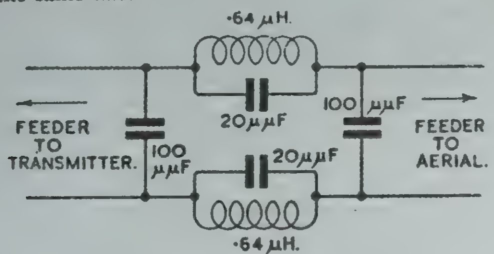


Fig. 16A.
Filter for use in 80–100 ohm balanced line between P.A. tank and aerial circuit or in low impedance feeder. Coils are 10 turns 16 S.W.G. on ½" former spaced to 1" winding length. Cut-off frequency about 18 Mc/s.

It should be pointed out that the harmonic suppression filters described have no effect at the fundamental frequency of the transmitter, having negligible loss at such a frequency. They will, however, attenuate very severely frequencies falling within the television range.

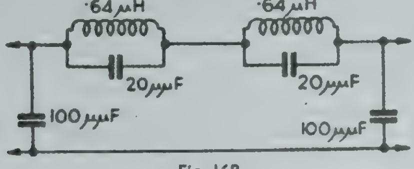


Fig. 16B.
Filter shown in Fig. 16A rearranged for use in 80–100 ohm concentric line. The centre conductor of the cable should be connected to the part of the filter having the inductances in circuit.

When used in feeders having an impedance of 300 or 600 ohms, slightly different values are required and these are given in Figs. 17 and 18. In Fig. 18B the same filter is slightly rearranged for use in a single wire feeder, such as a Windom.

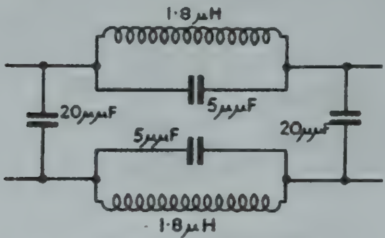


Fig. 17.
Rearrangement of filter shown in Fig. 16A for use in 300 ohm balanced line. Coils are 10 turns 18 S.W.G. on 1" dia. former spaced to 1" winding length.

The cut-off frequency of these circuits is about 18 Mc/s., although its exact value is immaterial. In each case an arbitrary cut-off has been chosen in order to quote condenser values which are conveniently obtainable.

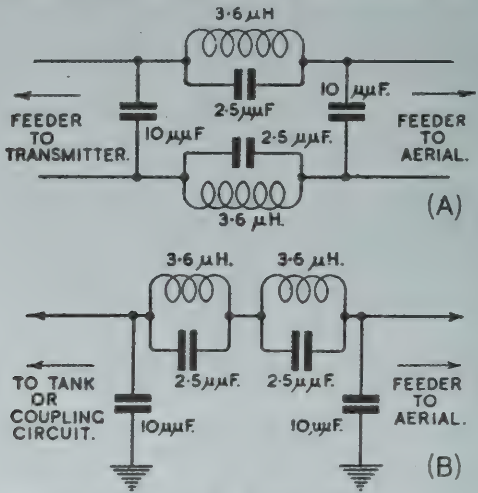


Fig. 18.
(A) Shows filter for use in 600 ohm balanced feeder.
(B) Re-arrangement of components for use in 600 ohms single wire feeder.
Coils are 18 turns 18 S.W.G. on 1" dia. former spaced to 2" winding length.

In these filters it is necessary to connect the condenser across the actual ends of the coil. Do not leave any length of lead on the coil. Also in the filter for the 80 ohm feeder keep the device as compact as possible, preferably soldering the condenser tags directly together without using any connecting wire. The feeder line need then be opened only an inch or so and connected

directly across the 100 $\mu\mu F$ condenser. For the 600 ohm circuit it will be necessary to place the coils at a distance apart corresponding to the spacing of the line. The leads to the 10 $\mu\mu F$ condenser should be straight, short lengths of stout wire or tape.

In cases where a matching network (Collins coupler) is used, the circuit can be modified to form an arrangement similar to the filters already described. This is done by connecting a small variable condenser across the series coils to tune them to the harmonic frequency, and will entail slightly resetting the other two condensers. Fig. 19 gives details of a suitable matching network.

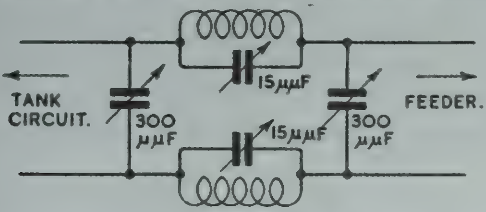


Fig. 19.
Suitable components for a matching network. Each coil to be made 5 turns on 2" diameter spaced to 1" winding length.

Where the optimum transmitter load and the feeder impedances are known, the circuit values can be calculated, but where an existing coupler is already working it should not be difficult to find the correct adjustments by trial and error.

It is advisable to use coils with no short-circuited turns. Such coils should be somewhat smaller in value than are normally required for 14 Mc/s. or difficulty may be met in tuning them up to 42 Mc/s., in fact the self-capacity alone may be sufficient, in which case resonance can be obtained at the correct frequency by varying the turns.

Using a normal type of coil with about five turns wound on a 2" former, a condenser with a maximum value of 15 $\mu\mu F$ should be suitable.

To adjust the circuit a harmonic indicator, as described on page 22, should be loosely coupled to the aerial side of the network by means of a very small capacity, say about 5 $\mu\mu F$. The condensers across the coupler coils should then be adjusted until minimum harmonic is observed. Both condensers should be kept at about the same setting until the tuning position is obtained, when individual fine adjustment can be made. The main condensers will now require readjustment to load the transmitter correctly, but this should not be appreciable if the circuit is functioning correctly.

If an unbalanced network is used, the adjustment will be simpler, as there is only one small condenser to adjust. A maximum capacity of 10 $\mu\mu F$ will probably be sufficient in this case as the coil will be about double the size used in the balanced variety.

If the separate tuned aerial tank circuit recommended above is used, it will be quite satisfactory to install one of these filters in the 80 ohm line coupling the P.A. to the tank circuit. In addition to the compact construction feature previously mentioned, the filter should not be coupled to the P.A. tank coil when fitted. The P.A. stage itself should, of course, be screened.

Harmonic Trap

Another very effective method of reducing harmonics is the use of a

harmonic trap in the anode circuit of the stage producing them. Normally the P.A. stage is the only one requiring this treatment. Fig. 20 shows how such a device should be fitted and gives the sizes of the coil and condenser needed. In the case of a push-pull P.A. stage, two traps will be required, one in each anode circuit.

The actual construction of the trap presents no difficulty. The condenser is mounted on a piece of Polystyrene about 2¼″ × 1¼″ × ⅛″ and supported on the anode tank condenser to allow very short connecting leads. The coil is soldered directly on to the trap condenser, with its axis at right angles to that of the tank coil. Any good quality midget air dielectric trimmer of capacity 3–20 μμF can be used. When calculating the size of the trap coil allowance must be made for the anode-to-filament capacity of the P.A. valve, and for stray capacities which may be effectively in parallel with the trap tuning capacity. The values given in Fig. 20 are for a valve having an anode-to-filament capacity of 0·4 μμF—such as a 100TH. The range of condenser values suggested will probably be suitable for most common types of transmitting valve.

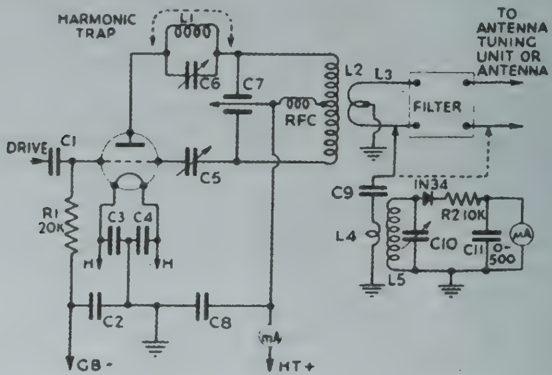


Fig. 20.
Typical P.A. circuit showing harmonic trap L1, C6, and harmonic indicator, together with aerial filter.

C1	50 μμF.	C11	·001 μF.
C2, 3, 4	Tubular paper. ·01 μF.	L1	12 turns 18 S.W.G. bare copper wire ½″ I.D. spacing 1/16″.
C5	Neutralising.	L2	Normal tank coil.
C6	3–20 μμF midget air-trimmer double-spaced.	L3	Link coil.
C7	50–50 μμF transmitting variable.	L4	One turn link adjustable.
C8	·001 μF H.V. mica.	L5	4 turns 14 S.W.G. enamelled copper 1½″ I.D. 1/16″ spacing.
C9	2–5 μμF.	R1	20,000 ohms 10 watt.
C10	7–100 μμF midget air-trimmer.	R2	10,000 ohms ½ watt.

Harmonic Indicator

It is essential for the proper adjustment of the trap that some form of indicator be used which is capable of being tuned to the harmonic it is desired to eliminate. A simple device on the lines of that indicated in Fig. 21, should be made up. As shown, this consists merely of a tuned circuit, which is loosely coupled *via* a small capacity to the transmitter, a crystal detector

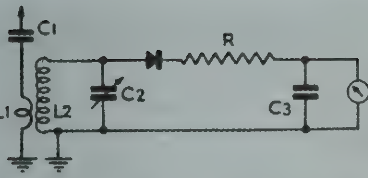


Fig. 21.
Circuit of Harmonic Indicator.
L1 1-turn link. L2 4 turns 14 S.W.G. $1\frac{1}{8}''$ diameter, spacing $\frac{1}{16}''$. C1 2–5 $\mu\mu$F Ceramic. C2 *Eddystone* type No. 583, with both sections in parallel. C3 ·002 or ·001 μF. R 10,000 ohms. M 0–500 microammeter, *Sangamo Weston*. Crystal detector *B.T.H.* type CG1C or *Silicon* type CS7A.

of the miniature micro wave type and a micro ammeter. Further minor refinements, such as connecting sockets and a phone jack, can be added if desired. The instrument described is illustrated in Fig. 22. The phone jack is included so that the indicator may be used as a monitor in the event of modulation of the transmitter causing spurious emissions on television frequencies. This is more likely to happen under conditions of over-modulation.

The indicator should be coupled to the output from the P.A. stage just tightly enough to give a full scale deflection on the meter when the circuit is tuned to resonance with the harmonic which is to be eliminated. It should be remembered that if a Silicon crystal is used it has very definite current limitations, and the meter needle should not be allowed to go "hard over" against the stop. A Germanium crystal on the other hand will pass more than sufficient current to burn out the meter. Equal care should therefore be taken if a crystal of this type is used.

The method of using the device is as follows. The transmitter is first tuned-up in the normal manner with the P.A. anode harmonic trap circuit (or circuits if push-pull is used), short-circuited temporarily by a short lead fitted with crocodile clips. The harmonic indicator is then coupled to one side of the feeder line (or output link coil) through a small capacity of about 2 to 5 $\mu\mu$F as shown in Fig. 20. The indicator is then tuned to resonance with the P.A. harmonic which falls in the television band. This will produce maximum deflection on the meter. The coupling to the indicator should now be adjusted so as to produce full-scale deflection of the meter under this

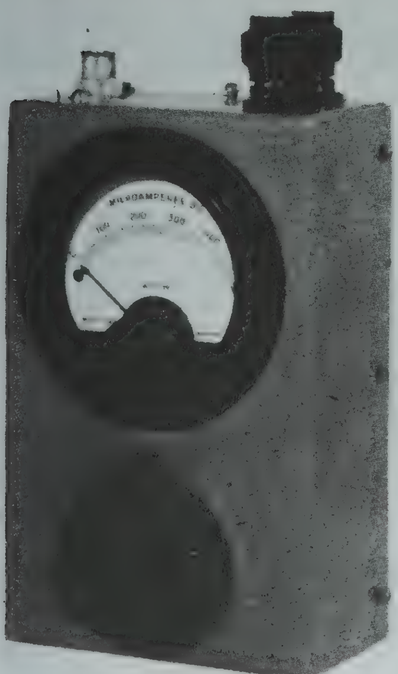


Fig. 22.
Harmonic Indicator. The miniature dial is an *Eddystone* type No. 595. The unit is contained in a metal box measuring $6'' \times 3\frac{3}{4} \times 2''$ obtainable from *Amateur Radio Service* (G6HP). A *Belling Lee* concentric socket and *Belling Lee* insulated terminals are provided for alternative connection to the transmitter.

reference condition. The harmonic trap shorting-link is now removed and the harmonic trap tuning control rotated by means of an extension spindle, until the indicator meter shows a minimum point.

In the case of a low power transmitter it may be necessary to connect the harmonic indicator link coil across the feeder through a small capacity in each leg—say 5 to 20 $\mu\mu$F—in order to obtain sufficient deflection of the microammeter. With large changes of frequency it will be necessary to readjust the trap tuning slightly, and in view of this it is strongly recommended that this control be brought out to the front panel for easy accessibility. In the same way a built-in harmonic indicator would be a great advantage, and its inclusion in the design of a 14 Mc/s. P.A. is greatly to be desired.

The indicator being very loosely coupled to the feeder produces negligible loading or unbalance. It may thus be left in circuit, when it will function as a permanent monitor of the transmitter harmonic output.

Should a feeder harmonic filter be used in addition to the anode trap circuit, its efficacy may be judged by shorting-out the trap circuit and observing the indicator deflection before the filter (*i.e.* full scale) and then clipping the indicator coupling condenser on to the feeder immediately after the filter and reading the indication thus produced. If necessary, the filter elements may be adjusted by this method for maximum rejection in the television band.

The 21 Mc/s. Band

When the 21 Mc/s. band is used trouble may be expected from 2nd harmonic radiation, as indicated in the table on page 16. The easiest method of reducing this source of interference either completely or to negligible proportions is to use a push-pull P.A. stage, since in a properly balanced circuit of this type the even harmonics cancel out. Care should, however, be taken to ensure that the stages are properly balanced, and since the two valves are unlikely to be exactly matched, the theoretically correct method of equalising the drive to each grid will not hold good in practice. The harmonic indicator should be coupled to the feeders *via* a very small capacity and the drive to each valve adjusted until the harmonic disappears. A similar result can be obtained by adjusting the bias on each valve individually, for example by using separate grid leaks or cathode resistors, and adjusting each until the 2nd harmonic is at a minimum.

Alternatively, if a single-ended P.A. stage is used, an anode trap similar to the one recommended and described for use with the 14 Mc/s. P.A. should be employed. The harmonic indicator will, of course, prove invaluable again in this case.

Modulators

Occasionally interference can be caused by Class B Modulators which emit spurious oscillations at R.F. on peaks of modulation only, and which modulate or radiate directly. The cure for this is to improve the modulator stability by using anode and/or grid stoppers, or any of the usual methods described either in the *Amateur Radio Handbook* or elsewhere.

Mains Filters

The design of a suitable mains filter is illustrated in Fig. 23. Such a device should be constructed in a metal box with a screen between the coils, and mounted in the main supply lead to the transmitter. A somewhat lighter construction can be used when it is intended to install the filter in the mains lead to the television receiver. Particular attention should be paid to the

insulation of the components. For this reason good quality condensers are essential. When a mains filter is fitted to a receiver the lead from the filter to the set should be short and screened.

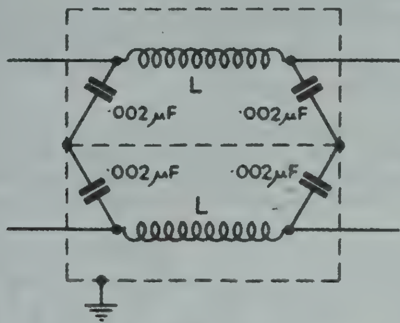


Fig. 23.
Circuit of mains filter to reduce 43 Mc/s. R.F. radiation. Dotted line shows screening box. The condensers should be mica 1,000v. test and of good quality. L = 69" of wire wound on 1" dia. former. For transmitter supply use 12 S.W.G., and for receiver supply 16 S.W.G.

High-Pass Filters

As stated previously, where interference is caused even when the amateur transmitter is radiating only in the licensed frequency bands, and where as much harmonic radiation as possible has been suppressed, a filter in the receiver aerial lead is usually the solution. Such a filter must pass television

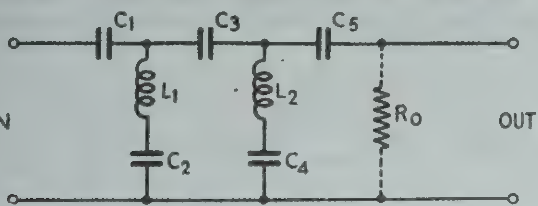


Fig. 24.
Simple two section filter with cut-off frequency abou: 30 Mc/s. and maximum attenuation at 14·3 Mc/s.

L1, L2	7 turns 16 S.W.G. on ⅜" dia. former spaced to ¾" winding length. Length of coil leads not to exceed ¾" at each end.
C1, C5	71 μμF T.C.C., type CM23N. C3 35 μμF T.C.C., type CM23N.
C2, C4	520 μμF (470 and 47 μμF, type CM24N and CM23N, in parallel).
R0	80 ohms ½ watt. (when required).

frequencies without loss, and severely attenuate all signals on lower frequencies.

The circuit and details of a simple M-derived high-pass filter are shown in Fig. 24. This design attenuates frequencies of the order of 30 Mc/s. and below, but has maximum loss at 14·3 Mc/s. [illegible] filter of this type should be constructed in a metal box with a screen betw[illegible] the two coils. The earthing of the screening box will be automatic i[illegible] [illegible]ncentric sockets are used for the input and output connections. The 80 o[illegible] terminating resistance (R), need only be used where the input impeda[illegible] of the receiver is

greatly in excess of this figure. The device should be included in the 80 ohms television receiver line as close to the receiver as possible.

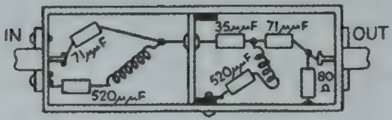


Fig. 25.
Layout of components of a simple 14 Mc/s. filter. The screening box measures 1½" deep, 2½" wide, and 5" long.

A further design of high-pass filter is shown in Fig. 26, but in this case maximum attenuation occurs at 29 Mc/s. Such a filter is intended mainly for use when interference is caused from a transmitter operated in the 28–30 Mc/s. band. It should be included as before in the receiver aerial feeder

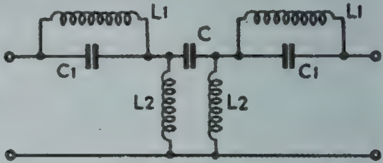


Fig. 26.
Circuit of high-pass filter with maximum attenuation at 29 Mc/s. C = 25 μμF. C1 = 100 μμF. L1 and L2 = ·285 μH = 4½ turns 18 S.W.G., wound 16 T.P.I. on ½" dia. former.

close to the set. Constructional details are given in Fig. 27. A filter of this type will cure trouble caused by 28 Mc/s. signals reaching the receiver through the aerial lead, but it will not help when the interference is mains-borne or picked up directly by the set wiring. As can be seen by a study of Fig. 28, the filter will attenuate signals on all frequencies lower than 29 Mc/s., in fact it will probably be found effective on 14 Mc/s. transmissions except where the interference is severe. In the latter event the filter shown in Fig. 24 should be used in addition.

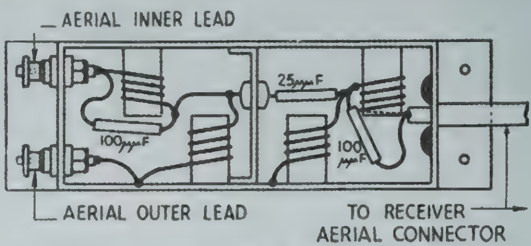


Fig. 27.
Layout of components of a 29 Mc/s. filter, in screening box. The braiding of the co-axial cable is unravelled and soldered to the inside of the box, unless concentric sockets are used. Terminals must be insulated from box, which measures 1½" deep, 2½" wide, and 5" long.

Single Side-Band Reception

A further me[illegible]od that has proved satisfactory in some cases, is to modify the receiver for single si[illegible] band reception using the upper side-band. This has the effect of redu[illegible]g the band-width of the overall response thereby reducing the noise [illegible]al and increasing the sound channel rejection. It should be rememb[illegible] that most of the 42 Mc/s. interference from amateurs falls in the lower [illegible]-band. For areas of very low television field-strength

the reduced overall noise level should be advantageous, quite apart from the considerable reduction in sensitivity in the region of the 3rd harmonic from a 14 Mc/s. transmitter. In addition, there will be appreciable rejection of the 12th harmonic of a 3·5 Mc/s. transmitter and the 6th harmonic of all 7 Mc/s. transmissions, all of which in low field-strength areas might have a very damaging effect upon the picture.

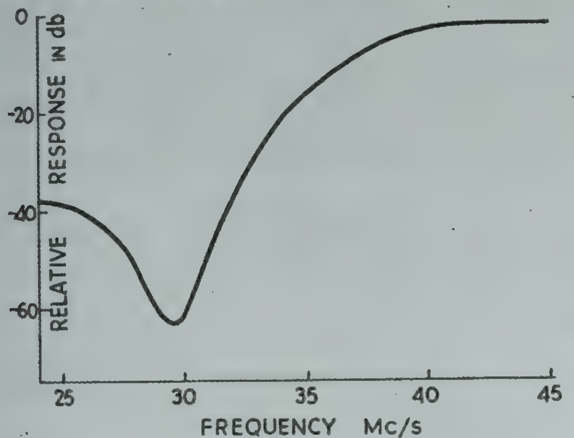


Fig. 28.
Response curve of 29 Mc/s. filter showing the points where maximum attenuation occurs.

The degree of rejection obtained by this means may be of the order of 25 db or more.

The conversion to single side-band reception should not be attempted by any other than an experienced television engineer, but the procedure is not difficult and good results can be obtained quite easily.

Obviously the owner must be willing to have the conversion made, and considerable tact should be used in suggesting any modifications to his apparatus. The general improvement in the signal-to-noise ratio, however, is an important point and should be explained carefully. Needless to say as much harmonic suppression as possible should already be in use in the transmitter.

Summary

From the foregoing it will be seen that the curing of Television interference from amateur radio transmitters, although at first sight apparently a difficult problem, is by no means insuperable. It is an aspect of Amateur Radio that is becoming rapidly more important with the increasing area of the Television service, coupled with longer programme hours. This, combined with the rapidly expanding number of viewers renders it an essential problem that must be solved by almost all amateurs sooner or later. Fortunately, in nearly every case, one of the remedies suggested, or a combination of two or more, will prove effective.

Once again it is urged that immediate attention be paid to any complaints received, and the matter investigated with as much tact, patience, and forebearance as the individual case requires.

Band. Mc/s.	Harmonic.	Range. Mc/s.	Chance of Interference.
1·715– 2·0	30th, 38th,	51·5–76	Negligible
3·5 – 3·8	16th, 17th, 18th	56 –68·4	Moderate
7 – 7·3	9th	63 –65·7	Slight
14·0 –14·4	4th	56 –57·6	Slight
21 –21·4	3rd	63 –64·2	Moderate
28 –30	2nd	56 –60	Serious
58·5 –60	Fundamental	58·5–60	Serious

TABLE III.

Shows the position for the frequencies to be used by the Birmingham Television Station.

Car Ignition

Although not strictly applicable to this booklet, a few points regarding the suppression of car ignition interference to television are given here so that amateurs may lead the way by suppressing their own vehicles, and be in a position to answer questions on the subject.

In general the greatest source of trouble is the ignition system, and the simplest cure in most cases is to insert a resistor having a value between 5,000 and 15,000 ohms in the lead from the distributor to the coil, at the distributor end. A suitable resistor is made by *Belling & Lee*, and is specially constructed so that it may easily be connected to the distributor and lead. The type No. is L1274, and the price is 1/6. If interference still persists then a resistor of any value between 5,000 and 25,000 ohms should be inserted at the plug end of each of the leads from the distributor to the sparking plugs.

If it is desired to suppress still further, another series of resistors may be inserted at the distributor end of these leads, provided the total resistance in each lead does not exceed 25,000 ohms.

Windscreen wipers are another source of trouble and may be suppressed by fixing a condenser of ·5 or 1 μF from each terminal to chassis. Electric petrol pumps may be suppressed in the same way.

In the case of the dynamo, if it is of the third brush type, which is more usual, condensers of ·5 or 1 μF should be connected between the field terminal and frame, and between the insulated brush terminal, or terminals, and the frame.

In every case connecting leads should be kept as short as possible and should not exceed 3" in length at any point.

In reply to any argument that suppression affects performance, it should be pointed out that all Service vehicles are suppressed to a very high degree, and after exhaustive bench and road tests no falling-off of performance is apparent. Furthermore, the Automobile Research Association has investigated the matter and concludes that no effect due to the use of correctly designed suppressors can be detected at various speeds and under load.

APPENDIX

This appendix gives a list of the various types of Television receivers in order of makers names and shows the type of circuit employed in each case. Where a superheterodyne circuit is used details of the intermediate frequency, oscillator and image frequency are given.

Except where otherwise stated the receivers utilise both sidebands. This implies, in the case of the vision portion of the circuit, that the effective intermediate and image frequency bandwiths extend about 2 to 2·5 megacycles either side of the figures quoted.

Name	Model	Circuit		I.F. in Mc/s.		Oscillator Frequency in Mc/s.	Image Frequency in Mc/s.	
		Vision	Sound	Vision	Sound		Vision	Sound
Alba ..	T411	T.R.F.	T.R.F.	—	—	—	—	—
	T421	T.R.F.	T.R.F.	—	—	—	—	—
Beethoven ..		S.H.	S.H.	13·0	16·5			
Bush ..	T91	T.R.F.	S.H.	—	·725	40·725	—	40·0
	TV1	T.R.F.	S.H.	—	·725	40·725	—	40·0
	TV2	T.R.F.	S.H.	—	·725	40·725	—	40·0
Ecko ..	All	T.R.F.	T.R.F.	—	—	—	—	—
Cossor ..	54	S.H.	S.H.	6·0	2·5	39·0	33·0	36·5
	65	S.H.	S.H.	6·0	2·5	39·0	33·0	36·5
	137 237	S.H.	S.H.	5·3	1·8	39·7	34·4	37·9
	437	S.H.	S.H.	5·3	1·8	39·7	34·4	37·9
	900 1200	S.H.	S.H.	6·0	2·5	39·0	33·0	36·5
	901, 902, 912	T.R.F.	S.H.	—	2·2	39·3	—	37·1
Dynatron ..	TV21	T.R.F.	T.R.F.	—	—	—	—	—
Etronic ..	T12	T.R.F.	T.R.F.	—	—	—	—	—
Ferguson ..	841T	T.R.F.	T.R.F.	—	—	—	—	—
	842T	T.R.F.	T.R.F.	—	—	—	—	—
Ferranti ..	T136	T.R.F.	S.H.	—	10·0	31·5	—	21·5
	T138	T.R.F.	T.R.F.	—	—	—	—	—
	T1146	T.R.F.	S.H.	—	10·0	31·5	—	21·5
	T1246	T.R.F.	S.H.	—	10·0	31·5	—	21·5
G.E.C. ..	BT0124	S.H.	S.H.	6	2·5	39·0	33·0	36·5
	BT3701	S.H.	S.H.	3	·45	41·95	38·95	42·4
	BT3702	S.H.	S.H.	3	·45	41·95	38·95	42·4
	BT8090	S.H.	S.H.	3	·6	42·1	39·1	42·7
	BT8121	S.H.	S.H.	3	·45	41·95	38·95	42·4
	BT8161	S.H.	S.H.	3	·45	41·95	38·95	42·4
	BT9121	S.H.	S.H.	6	2·5	39·0	33·0	36·5

Name	Model	Circuit		I.F. in Mc/s.		Oscillator Frequency in Mc/s.	Image Frequency in Mc/s.	
		Vision	Sound	Vision	Sound		Vision	Sound
G.E.C. ..	BT9122	S.H.	S.H.	3	·45	41·95	38·95	42·4
	BT7092	S.H.	S.H.	13·5	10·0	31·5	18·0	21·5
	BT7094	S.H.	S.H.	13·5	10·0	31·5	18·0	21·5
H.M.V. ..	900	T.R.F.	S.H.	—	·46	41·04	—	40·58
	901	T.R.F.	S.H.	—	1·5	40·0	—	38·5
	902	T.R.F.	S.H.	—	·46	41·04	—	40·58
	902A	T.R.F.	S.H.	—	·46	41·04	—	40·58
	903	S.H.	S.H.	9·0	5·5	36·0	27·0	30·5
	904	S.H.	S.H.	8·0	4·5	37·0	29·0	32·5
	905	S.H.	S.H.	8·0	4·5	37·0	29·0	32·5
	907	S.H.	S.H.	8·0	4·5	37·0	29·0	32·5
	1800	S.H.	S.H.	8·0	4·5	37·0	29·0	32·5
	1801	S.H.	S.H.	8·0	4·5	37·0	29·0	32·5
	1802	S.H.	S.H.	8·0	4·5	37·0	29·0	32·5
	1803	S.H.	S.H.	8·0	4·5	37·0	29·0	32·5
	1804	S.H.	S.H.	8·0	4·5	37·0	29·0	32·5
	1850	S.H.	S.H.	8·0	4·5	37·0	29·0	32·5
	1901	S.H.	S.H.	8·0	4·5	37·0	29·0	32·5
Haynes ..	HR77	T.R.F.	S.H.	—	5·0	36·5	—	31·5
	HR88	T.R.F.	S.H.	—	3·0	38·5	—	35·5
Invicta ..	All	T.R.F.	T.R.F.	—	—	—	—	—
K.B. ..		S.H.	S.H.	12·5†	9·0	32·5	20·0	23·5
McMichael	De Luxe	S.H.	S.H.	13·0	9·5	32·0	19·0	22·5
	Standard	T.R.F.	T.R.F.	—	—	—	—	—
Marconiphone ..	701	T.R.F.	S.H.	—	·46	41·04	—	40·58
	702	T.R.F.	S.H.	—	1·5	40·0	—	38·5
	703	T.R.F.	S.H.	—	·46	41·04	—	40·58
	704	S.H.	S.H.	9	5·5	36·0	27·0	30·5
	705	T.R.F.	S.H.	—	·46	41·04	—	40·58
	706	S.H.	S.H.	8·0	4·5	37·0	29·0	32·5
	707	S.H.	S.H.	8·0	4·5	37·0	29·0	32·5
	709	S.H.	S.H.	8·0	4·5	37·0	29·0	32·5
	710	S.H.	S.H.	8·0	4·5	37·0	29·0	32·5
	711	S.H.	S.H.	8·0	4·5	37·0	29·0	32·5
	712	S.H.	S.H.	8·0	4·5	37·0	29·0	32·5
	713	S.H.	S.H.	8·0	4·5	37·0	29·0	32·5
	VT50A	S.H.	S.H.	8·0	4·5	37·0	29·0	32·5
Murphy ..	A42V	S.H.	S.H.	4·25	·75	40·75	36·5	40·0
	A56V	S.H.	S.H.	4·25	·75	40·75	36·5	40·0
	A58V	S.H.	S.H.	4·25	·75	40·75	36·5	40·0
	V86CA	S.H.	S.H.	13·5	10·0	31·5	18·0	21·5
	V114	S.H.	S.H.	13·5	10·0	31·5	18·0	21·5
	V116	S.H.	S.H.	13·5	10·0	31·5	18·0	21·5
	V134	T.R.F†	T.R.F.	—	—	—	—	—
	V136	T.R.F†	T.R.F.	—	—	—	—	—

Name	Model	Circuit		I.F. in Mc/s.		Oscillator Frequency in Mc/s.	Image Frequency in Mc/s.	
		Vision	Sound	Vision	Sound		Vision	Sound
Philco ..	A1707	T.R.F.	T.R.F.	—	—	—	—	—
	A1708	T.R.F.	T.R.F.	—	—	—	—	—
Philips ..	All	S.H.	S.H.	13·2*	9·7	31·8	18·6	22·1
Pilot ..	VS9	T.R.F.	T.R.F.	—	—	—	—	—
Pye ..	B16T	T.R.F.	T.R.F.	—	—	—	—	—
	D16T	T.R.F.	T.R.F.	—	—	—	—	—
	B18T	T.R.F.	T.R.F.	—	—	—	—	—
	D18T	T.R.F.	T.R.F.	—	—	—	—	—
R.G.D. ..	All	S.H.	S.H.	13·0	9·5	32·0	19·0	22·5
Ultra ..	T22	S.H.	S.H.	5·8 *	2·3	39·2	33·4	36·9
	V470	S.H.	S.H.	10·7 †	7·2	34·3	23·6	27·1

* Single Side Band (Lower)
† Single Side Band (Upper)

Acknowledgements

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CONTENTS

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FOREWORD

TO cover the subject of amateur receiving equipment adequately in every detail would occupy far more space than is available in this small book, notwithstanding the restriction of the present treatment to the reception of frequencies below 30 Mc/s. During the last twenty years intensive research has led to highly elaborated circuit techniques, especially in the superhet type of receiver, and numerous journals and textbooks have contained specialised descriptions of the advances that have contributed to the present-day methods and designs. The reader may find it necessary to refer to such other publications for specific details of various matters not included here, but it is hoped that this book will assist in guiding him through the intricate mass of technicalities and make it easier for him to assimilate the more highly specialised knowledge contained in those publications.

While care has been taken to avoid the use of workshop jargon which is so frequently encountered in the practice of Amateur Radio, no apology is offered for the use of the word *superhet*. Originally the receiver which is now known in many different languages as a *superhet* was called a *supersonic heterodyne* receiver. Within a few years of its conception, this term was shortened to *superheterodyne*, but due partly to the difficulties experienced in various tongues and partly to the constant desire for abbreviation, the word *superhet* now seems to have become universally established.

Other awkward phrases, such as *D.C. current*, *A.C. voltage* or *I.F. frequency*, are common enough and perhaps tolerable in everyday speech, but an attempt has been made here to evade them by using the rather more rigorous style permitted in technical writing.

Numerous references are made in the text to two other publications in the same series as this book: *V.H.F. Technique* and *Valve Technique*. Some overlapping of subject-matter has been inevitable in the discussions of fundamental principles, but the reader would be well advised to regard them as companions to this book.

The author is indebted to Mr. E. L. Gardiner, B.Sc., for his suggestions concerning the material to be included here and its arrangement. Acknowledgement is also gratefully made to *Messrs. Denco Ltd.* for the use of photographs illustrating various constructional features.

CHAPTER 1 FUNDAMENTAL PRINCIPLES

THE two systems commonly used in Amateur Radio communication are telegraphy and telephony. Generally speaking, a receiver can be made suitable for both systems, but an instrument which is designed especially for one is not necessarily suitable for the other.

In radio telegraphy the system in widest use in known as *Continuous Wave*, abbreviated to *C.W.* A generator of continuous waves (otherwise describable as a *C.W. Oscillator*) supplies an alternating current of steady (*i.e.* continuous) amplitude. Since the transmission is to be radiated through the ether, the frequency of this alternating current will be at least 10 kc/s. and may be much higher—possibly of the order of 10 Mc/s. or even 1,000 Mc/s. The practical limit at the present time is considered to be about 10,000 Mc/s.

In radio telephony, the speech or music is used to modulate the C.W. oscillation, which is then referred to as the *carrier*. Various systems of modulation are in use, *e.g.* amplitude, frequency and pulse modulation, and in each case the receiver must have characteristics which are especially suited to the system.

Unmodulated C.W.

In the simplest form of radio telegraphy transmission a key is inserted in the transmitter circuit so that the C.W. oscillations are radiated only as required in accordance with the Morse code. Each dot or dash thus consists of a train of uniform oscillations. No direct use can be made of these oscillations when they reach the receiver and it is necessary to convert the incoming signals into a form which can be perceived by the human senses. The usual practice is to convert the frequency of the oscillations to a much lower frequency—so much lower, in fact, that it becomes audible. This process is known as *heterodyning*.

Modulated C.W. (Amplitude Modulation)

The amplitude of a C.W. oscillation can be varied, if necessary, in a very complex manner such as is found in speech and music. If this is done, the signal can be made audible merely by including in the receiver any device which responds to the changes in amplitude.

Fig. 1 shows an oscillation of varying amplitude. It is important to note

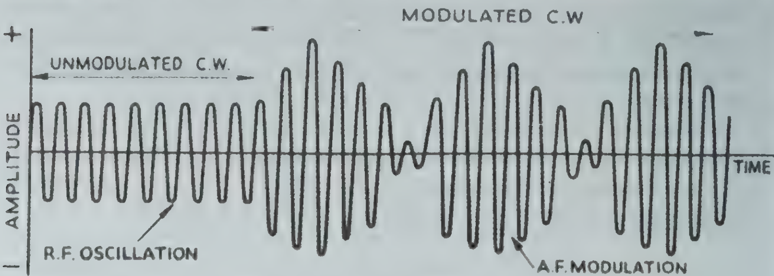


Fig. 1.
A continuous-wave carrier showing the form of modulation that occurs during the transmission of speech or music.

that the mean frequency remains constant. When the amplitude varies there is inevitably a change in the character of each alternation, and this has the effect of introducing certain other frequency components. A mathematical analysis shows that the modulated carrier may be considered to consist of an unmodulated carrier and a varying number of closely associated frequencies within the so-called side-bands.

If the incoming modulated signal is rectified, the waveform, Fig. 2A, will be bisected as shown in Fig. 2B. When suitably filtered so as to smooth out all the individual pulsations, the signal acquires the waveform shown in Fig. 2C. This is what is known as *detection* or *demodulation*. An undulation of this kind, consisting of various frequencies within the audio range, can be made audible by means of headphones or a loudspeaker: the output will be of the form shown in Fig. 2D.

These relatively low-frequency variations, which existed as variations in the amplitude of the radio-frequency current, were, so to speak, carried by the radio-frequency current, and the latter is therefore known as the *carrier*. The low-frequency (or audio-frequency) variations are known as the *modulation* of the carrier. The system is referred to as *amplitude modulation*.

A modulation consisting of a steady tone (*i.e.* constant frequency and amplitude) may be "keyed" to provide Modulated C.W. (*M.C.W.*) telegraphy. The carrier may be keyed together with the modulation or left running continuously, irrespective of the keyed tone.

Modulated C.W. (Frequency Modulation)

Instead of varying the amplitude of the carrier as the means of modulation, it is sometimes preferable to vary the frequency: if frequency variation is used, it is desirable to keep the amplitude constant. The system is therefore the converse of amplitude modulation, thus:

System		*Amplitude*		*Frequency*
Amplitude Modulation	..	Varies	..	Constant
Frequency Modulation	..	Constant	..	Varies

In a frequency-modulation receiver, the circuit is designed to be unresponsive to changes of amplitude and to be responsive only to changes of frequency. The *extent* to which the frequency is deviated above and below the mean frequency depends on the instantaneous *strength* of the modulating signal in the transmitter: the *rate* at which it is varied depends on the *frequency* of the modulating signal.

In order to transmit a strongly modulated signal, and thereby to produce a high signal/noise ratio at the receiver, it is necessary to use a relatively wide frequency deviation. The bandwidth required for an F.M. system is therefore large compared with the corresponding bandwidth required for an A.M. system, and for this reason it is usual to restrict F.M. transmissions to the higher frequency ranges in the communications spectrum. Frequency modulation is, in fact, best associated with V.H.F. operation, and the reader who is specially interested in the subject is referred to *V.H.F. Technique*, another booklet in this series.

Phase Modulation

For the sake of completeness, some mention should be made of *phase modulation*. In this system, the modulating signal in the transmitter is made to shift the phase of the carrier forwards and backwards in relation to the unmodulated condition. This has the effect of varying the frequency above

and below the normal value and to all intents and purposes, at least at the receiving end, the system is indistinguishable from frequency modulation, and it is similarly best restricted to V.H.F. channels.

Pulse Modulation

In recent years there have been important developments in yet another system, known as *pulse modulation*, but so far its usefulness has been limited to commercial purposes. The bandwidth required for pulse modulation is very large, and this system is tolerable only in the V.H.F. bands.

Amplification

Except in very special circumstances, the amount of energy induced in a receiving aerial is extremely small. If the power of the transmitter is relatively high and the distance from the transmitter is sufficiently short, the energy in the receiving aerial can be used directly—as in the old-fashioned crystal receiver—to actuate the headphones without amplification. In this case, the sound energy by which the operator is made aware of the signal has been supplied by the transmitter.

It is almost universal practice nowadays to amplify the incoming signal by means of valve amplifiers: this enables the receiver to respond to much weaker signals and permits the use of loudspeakers. The energy in the sound waves produced by a loudspeaker may be many million times greater than the energy in the radiation received by the aerial.

The study of amplification is now very highly developed and is of vital importance not only in the design of receivers but in transmitters and auxiliary equipment for measuring and test purposes. A detailed treatment of the methods by which a valve can be made to amplify is given in *Valve Technique*, another booklet in this series. It will be sufficient to note that in the orthodox amplifier, the signal to be amplified is fed to the circuit connecting the grid and cathode of the valve, while the output circuit is connected in series

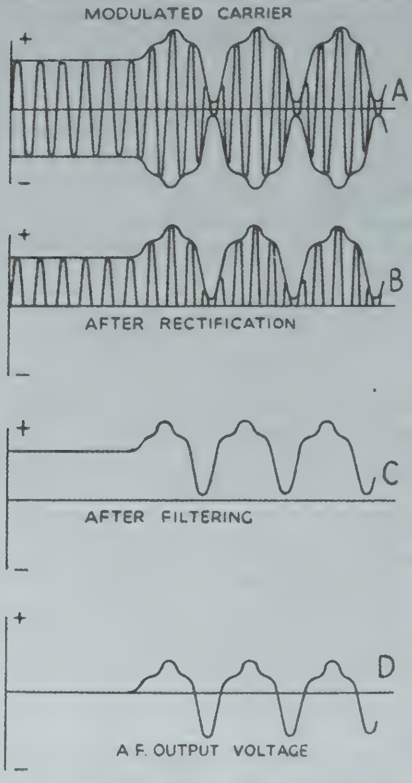


Fig. 2.
The detection or demodulation of a modulated carrier, showing how the A.F. signal is derived.

with the cathode-anode path in the valve and to the source of energy. The changes in grid potential (relative to the cathode, which is normally kept at a steady potential) cause changes in the electron current flowing through the valve and thus in the current flowing in the output circuit.

Amplifying valves may have only three electrodes (as in the single-grid triode) or several more. Most valves designed for amplification are of the screen-grid type (either tetrodes or pentodes). The purpose of the additional electrodes is to improve the sensitivity and power-handling capacity and to reduce harmful effects such as self-oscillation and distortion.

A considerable degree of amplification can be achieved by one valve alone, but two or more valves in succession are often used to obtain greater amplification. The total amplification in a multi-stage amplifier is, of course, the *product* of the various degrees of amplification provided by each of the successive stages. Multi-stage amplifiers are usually constructed on a metal chassis with thorough screening of the input circuits. These features are important, for otherwise the performance of the amplifier will suffer by reason of unwanted electrical feedback or pickup.

Heterodyne Reception

It has already been stated that a C.W. transmission is not audible until the carrier frequency has been converted to a much lower frequency—within the audio range. This is achieved by the process of heterodyning the carrier. Suppose that the carrier frequency is 7 Mc/s. and suppose that another radio-frequency generator is made to supply a frequency of 7·001 Mc/s. If these two frequencies are fed to the grid of a valve having a non-linear (*i.e.* rectifying) characteristic, the anode current will vary with a very complex frequency. The several components will include the two input frequencies (7·0 Mc/s. and 7·001 Mc/s.), a frequency equal to the difference between them (0·001 Mc/s.), another equal to their sum (14·001 Mc/s.), and a series of others made up of the sum or difference of the various harmonics. The most important of these components is the difference between the two main frequencies. In this example, this is 0·001 Mc/s., *i.e.* 1,000 c/s., which is within the audio range. Assuming that the anode circuit of the valve is designed to respond to audio frequencies, the incoming signal at 7 Mc/s. will be heard as long as the auxiliary frequency of 7·001 Mc/s. is present. If the auxiliary frequency is *not* present there will be no sound. Neither will there by any sound if only the auxiliary frequency is present—*i.e.* without an incoming signal.

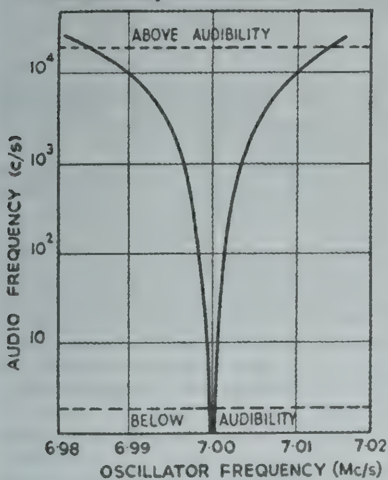


Fig. 3.
The variation of pitch of the beat note with oscillator frequency. At the silent centre, or zero-beat position, the oscillator frequency is equal to the carrier frequency.

If the 7 Mc/s. carrier comes from a C.W. transmitter which is radiating a Morse transmission, the operator at the receiver will then hear the Morse signals and the signals will have a pitch of 1,000 c/s. If he alters the frequency

of his auxiliary oscillator the pitch will also change, since the difference-frequency changes. Suppose that the frequency is reduced gradually so that it approaches the incoming carrier frequency, 7 Mc/s. The difference will steadily fall and the pitch of the note will therefore steadily fall. Eventually it will cease to be audible, when the difference becomes less than about 20 c/s. Suppose, further, that the operator continues to reduce the oscillator frequency. As it falls below the carrier frequency, the difference will now begin to increase. When it reaches 6·999 Mc/s., the note will again have a frequency of 1,000 c/s. If the difference is increased (either above or below) the pitch will, of course, increase and eventually it will become so high as to be inaudible.

Thus the *beat note*, as it is called, has the same frequency at two different tuning adjustments, equally spaced on either side of the centre, or zero-beat, position: see Fig. 3.

This heterodyne method of reception of C.W. signals is used in both the straight and the superhet type of receiver.

If the transmission consists of speech-modulated C.W., the presence of an auxiliary oscillation, which produces the beat note required for C.W. reception, would interfere very seriously with the reception of the speech. The local oscillator must therefore be switched off, or rendered ineffective, for telephony reception.

CHAPTER 2 PRINCIPLES OF RECEIVER DESIGN

THE essential parts of a simple receiver must serve the following purposes:

(*a*) To select signals at the specified frequency and reject all others.

(*b*) To demodulate, *i.e.* detect, the selected signals.

(*c*) To make the signals audible.

These three functions are indicated in the block diagram, Fig. 4. Amplifiers may be added to this simple receiver to improve its sensitivity and to produce louder signals. An amplifier placed between the detector and the headphones must operate at A.F. If it is placed between the tuner and the detector it must operate at R.F. If it is placed between the aerial and the tuner it may be of the untuned (aperiodic) R.F. type or the tunable type (in which case it will be provided with its own tuned circuit). Fig. 5 is a block diagram indicating the positions of the three types of amplifier.

Any device having a greater conductivity in one direction than the other (assuming that it has very small self-capacitance) will serve to rectify R.F. currents and thus act as a detector. The simplest of these devices is the crystal detector.

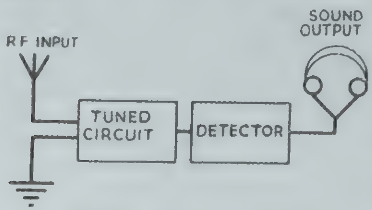


Fig. 4.
The essential functions of a receiver.

Crystal Detector

In Fig. 6A the tuned circuit is fed with A.C. applied to *X* and *Y*, and when it is adjusted to resonance an alternating voltage appears across the tuned circuit. On account of the unilateral conductivity of the crystal detector *D*, a unidirectional current flows through the resistance *R*. The condenser *C* serves to store the

energy and thus filter out the interruptions in the current which occur at every half-cycle.

In a simple crystal receiver, the resistance *R* is replaced by the headphones, and *X* and *Y* can be connected to the aerial and earth. For various reasons which are explained more fully in Chapter 3, it is better to connect the aerial to a separate inductance coupled to the main tuning circuit: see Fig. 6B. In this way the sensitivity and the selectivity are improved.

Diode Detector

A two-electrode valve (commonly called a *diode*) could equally well be used in place of the crystal detector: see Fig. 6C. Its sensitivity is of the same order as that of the crystal. The diode valve is very often used in receivers of the superhet type and for auxiliary purposes such as automatic volume control (A.V.C.) and noise suppression.

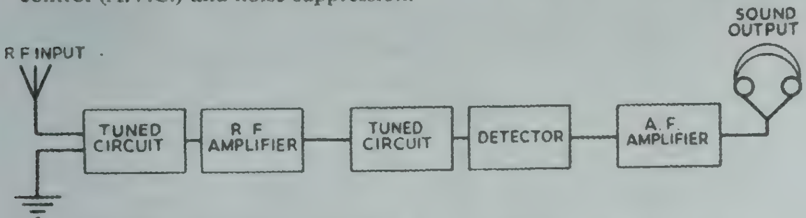


Fig. 5.
The use of amplifiers in a receiver at R.F. and A.F. To obtain improved selectivity the R.F. amplifier is preferably tunable, as shown, but for simplicity in construction and operation it may be aperiodic.

Triode Detector

An improved sensitivity can be achieved by using the amplifying properties of the triode valve and at the same time using the rectifying property due to the curvature of its characteristic curve. A typical curve is shown in Fig. 7. From this it is seen that the nature of the changes in the anode current depends on the value of the mean grid potential (*i.e.* the grid bias).

If the bias is adjusted so that the operating point is near the centre of the characteristic curve, the resultant anode-current change is an oscillation of substantially the same form as the oscillation applied to the grid, and the valve then acts merely as an amplifier. If, however, the grid bias is made suitably more negative so that the operating point occurs near the lower bend of the curve, the consequent effect on the change of anode current is to make the upward swing much greater than the downward swing: *i.e.* rectification is taking place.

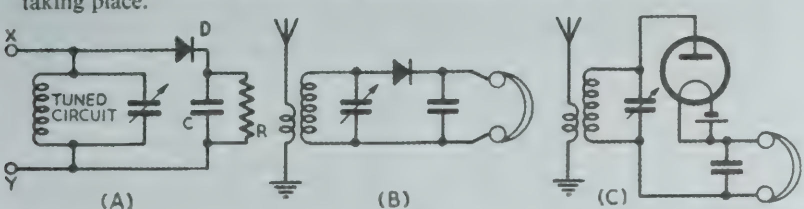


Fig. 6.
The radio-frequency voltage developed across a tuned circuit can be rectified to produce D.C. as in (A) or to produce audible signals in (B) and (C) if the R.F. voltage is modulated at audio frequency.

A similar rectification, although in the reverse sense, occurs if the operating point is chosen to lie near the upper bend of the characteristic, but this condition has the disadvantage of a higher mean value of anode current.

The efficiency of rectification in these examples depends on the difference in the slope of the curve on the opposite sides of the operating point, *i.e.* on the sharpness of the curvature. The system is known as *anode bend rectification.*

A suitable circuit using this form of detector is shown in Fig. 8. Since this arrangement is incapable of receiving C.W., it is of negligible value in communication practice, but it constitutes a useful monitor for telephony transmissions.

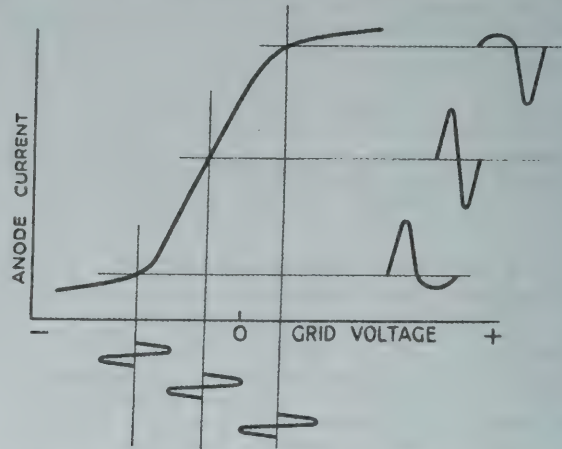


Fig. 7.
The characteristic curve of a triode has an upper and a lower bend, either of which produces a rectifying effect on oscillations applied to the grid. No rectification occurs if the oscillation is confined to the straight portion of the characteristic.

A further increase in sensitivity is possible with the same triode valve by using the grid as the equivalent of the anode of a diode detector arrangement and at the same time employing it to control the electron current flowing to the main anode in the manner of an ordinary triode amplifier. In Fig. 9 these two elementary circuits are shown side by side. Fig. 9A illustrates the diode detector arrangement (similar in principle to Fig. 6C) while Fig. 9B represents a simple triode amplifier circuit. By connecting *P—P* and *Q—Q*, a

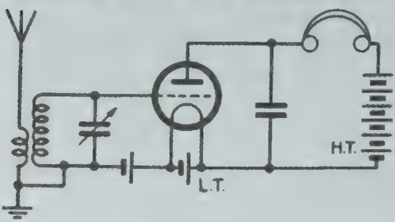


Fig. 8.
An elementary form of receiver using an anode-bend detector. It is suitable only for modulated signals and will not receive C.W. transmissions.

composite detector-amplifier is obtained. The circuit now takes the form shown in Fig. 10, which will be recognised as the popular *leaky-grid detector*.

The grid bias for the amplifier portion of this circuit is produced by the rectified grid current flowing through the grid leak R, but since the potential of the grid (*i.e.* the " diode " anode) cannot fall much below some small positive value—otherwise the input circuit would be blocked—the characteristics of the amplifier are thereby incapable of giving good quality of speech signals. Such a detector is likely to be overloaded by signals of quite moderate amplitude.

The two functions of diode rectification and triode amplification can be performed separately but combined in a single valve of the diode-triode type (*e.g.* DH63, 6SQ7): see Fig. 11. Here the triode grid receives its proper bias from the voltage-drop across the resistance in the cathode lead, and the output from the diode is taken from the points AB through the condenser C. This arrangement is very popular in superhets, but is not often used in straight receivers owing to the impossibility of providing R.F. feedback, *i.e.* the impossibility of producing self-oscillation of the detector.

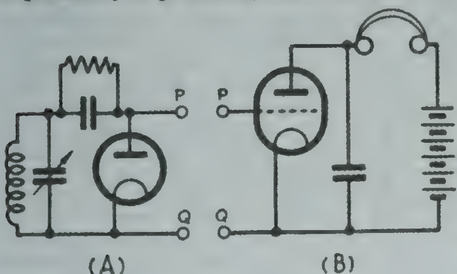


Fig. 9.
An analytical dissection of the leaky-grid detector into its constituents, *viz.*, a diode detector (A) and a triode amplifier (B).

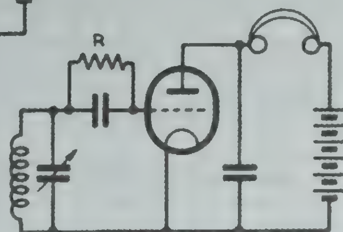


Fig. 10.
The leaky-grid detector in its usual form. The grid serves as the control electrode of the triode amplifier and as the anode of a diode rectifier.

Infinite-impedance Detector

A form of detector which gives efficient demodulation with very little distortion (less than the diode or the various forms of triode detector already described) is the so-called *infinite-impedance detector*. It is sometimes referred to as the *negative-feedback detector*. In this arrangement, the load in the anode circuit is so placed that it is also part of the grid circuit, thus providing 100 per cent. negative feedback: see Fig. 12.

The voltage developed across R when a modulated carrier is fed into this circuit contains R.F. components as well as A.F. components, and the R.F. is prevented from entering the headphones by the inclusion of the by-pass condenser C and the R.F. choke L. A blocking condenser is connected in series with the headphones to prevent the modification of the bias resistance R.

The input side of the valve behaves as an infinite impedance (by reason of the 100 per cent. negative feedback), and the circuit is therefore advantageous where it is necessary to maintain a high degree of selectivity in the tuned circuits. The output side has a very low resistance: this is also an advantage in certain circuits. For further details, see *Valve Technique*, pages 51–55.

Regenerative Detector

In a detector circuit, such as that shown in Fig. 8 or Fig. 10, the anode-current variations produced by an incoming signal have an R.F. component. Normally this is by-passed to earth by the condenser *C*. If an inductance is connected in the anode circuit, as shown by L_0 in Fig. 13, this R.F. current will create an R.F. magnetic field in and around the inductance coil. When this coil is coupled to the coil L_1 in the grid circuit, R.F. energy will be added to the grid circuit, provided that the sense of the coupling is such that the feedback is positive. This is called *regeneration*. If the coupling is reversed, the energy in the grid circuit will be diminished.

If the regeneration is increased sufficiently the amount of energy fed back to the grid will be greater than the amount of energy originally applied to it, and when this condition exists the circuit will produce continuous oscillations. The frequency is, of course, determined by the tuned circuit.

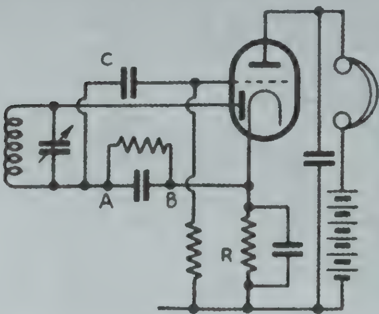


Fig. 11.
A popular detector-amplifier arrangement using a diode-triode valve.

A self-oscillating detector will heterodyne an incoming carrier, thus generating an audible beat note when the frequency-difference is small (up to about 10 kc s.), and if the carrier is modulated with speech signals, the reception of the speech will be completely spoiled by the heterodyne beat note. For receiving telephony, the feedback must therefore be reduced until self-oscillation ceases. If the feedback is further reduced below this point, the amplification diminishes, and it is usually preferable to keep the amount of regeneration as high as possible without actually causing self-oscillation. For C.W. reception, the detector is allowed to oscillate in order to produce an audible beat note: this is known as *autodyne reception.*

Generally it is found that the most suiable type of valve for a detector is a triode, but with a little care in circuit adjustment a tetrode or a pentode can be used and a rather higher sensitivity may then be expected.

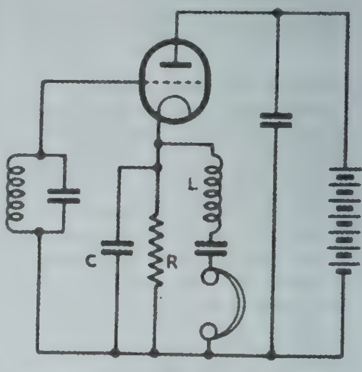


Fig. 12.
Infinite-impedance detector. The grid circuit receives 100 per cent. negative feedback from the anode circuit.

Super-regenerative Detector

A very interesting modification of the simple regenerative type of detector is the *super-regenerative detector*. Briefly the principle by which it operates is as follows. Strong feedback is used in the

detector to boost the signal strength, but the self-oscillation is interrupted at a supersonic frequency whereby the effect of the heterodyning of the carrier is rendered inaudible. The frequency of the interruption (known as the *quench frequency*) should be as low as possible in order to allow sufficient time in between the successive quenchings for the oscillations to build-up in magnitude, but if it is too low an unpleasant whistle will be superimposed on the signal.

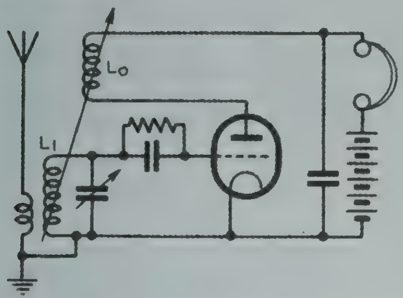


Fig. 13.
A simple regenerative detector circuit in which self-oscillation is produced as a result of positive feedback from the reaction coil L_0 to the signal tuning L_1. *N.B.*—This circuit radiates R.F. energy and may interfere with other receivers in the vicinity.

A receiver incorporating a super-regenerative detector is extremely sensitive but its selectivity is very poor. Where the bands are crowded with incoming signals this type of receiver is useless without some means of improving the selectivity. There is considerable distortion of speech and music, but the reproduction is usually comparable with "commercial" quality.

The effectiveness of the quenching system in boosting the signal strength depends on the carrier frequency and the quench frequency. The latter is usually not less than about 20 kc/s. and it is desirable not to make it higher than about 100 kc/s. The sensitivity increases as the carrier frequency increases, and the system is therefore more effective at V.H.F. than on the lower frequency bands.

For further details of the super-regenerative receiver, the reader is referred to *V.H.F. Technique*, pages 41–43.

Detector followed by A.F. Amplifier

Any A.F. amplifier may be used to boost the output from a detector, whether it be a crystal or a valve detector. The coupling between the detector and the amplifier should be designed to suit the respective impedances in order to ensure the maximum transfer of energy.

An A.F. transformer having a step-up ratio of anything between 2 : 1 and 8 : 1 is suitable for coupling an amplifier to a crystal or a diode or a triode detector, the higher ratios being preferable with a crystal. If the detector is a tetrode or a pentode, it would be necessary to use a transformer with a primary winding of very high impedance, and in these cases it is more convenient to use resistance-capacitance coupling.

Typical circuits showing the coupling arrangements for triode and tetrode detectors are given in Fig. 14. Usually a single stage of A.F. amplification is found sufficient for headphone reception, but two stages are sometimes preferable for loudspeaker operation.

For further details of A.F. Amplifiers the reader is referred to *Valve Technique*, pages 23–30.

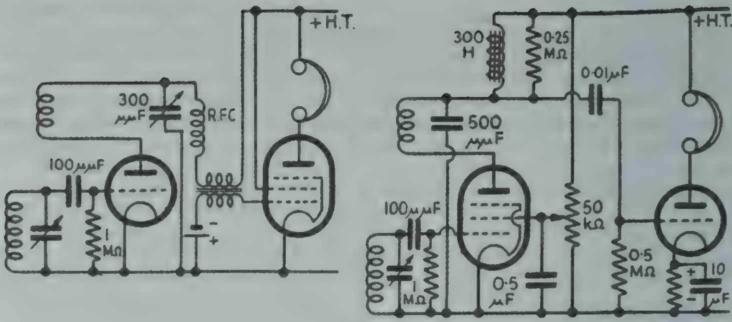


Fig. 14.
Typical coupling arrangements for triode and tetrode detectors followed by an A.F. amplifier.

Detector preceded by R.F. Amplifier

There are several advantages in the use of an R.F. amplifier. The amplitude of the incoming signals can be increased quite considerably, and the selectivity is also greatly improved. Further, the self-oscillating detector is isolated from the aerial circuit and there is consequently much less risk of causing interference to nearby receivers. Another very important advantage is the elimination of the effect which the aerial circuit would otherwise have on the tuning of the detector circuit: accurate calibration of the detector tuning thus becomes possible.

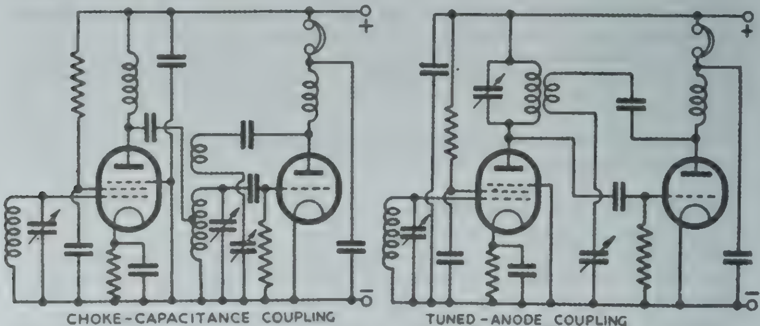


Fig. 15.
Typical coupling arrangements for a regenerative triode detector preceded by an R.F. amplifier.

Fig. 15 shows two typical circuits illustrating the coupling of an R.F. amplifier to a detector. It will be seen that in each arrangement there are two tuned circuits. These two circuits must both be tuned to the frequency of the incoming signals, and it is obviously preferable to link the two controls together. If the inductances are carefully designed, it is quite practicable to use a twin-gang condenser in such circuits.

One of the tunable circuits could be made aperiodic, but the advantage of the improved selectivity would then be lost and there would be a risk of cross-modulation by strong signals on an adjacent frequency.

Two stages of R.F. amplification can be used with advantage, but some difficulty may be experienced in using three stages, especially if maximum gain is required.

The Straight Receiver

A receiver in which the incoming signals are amplified directly, *i.e.* at their original frequency, and subsequently fed to a detector (which therefore also operates at the original frequency), is known as a *straight* receiver. The circuits given in Fig. 15 are covered by this description. The word " straight " may be taken to signify that all the tuning circuits must be adjusted to have the same frequency whereby all the tuning " points," so to speak, would be in a straight line. This type of receiver is also known as a *T.R.F.* receiver, *T.R.F.* meaning *tuned radio frequency* and obviously referring to the R.F. amplifier.

The straight receiver often comprises one R.F. stage preceding the detector and one A.F. stage following the detector, although for headphone reception the A.F. amplifier is not always necessary. To some extent, the A.F. amplifier may be considered to compensate for the not-very-high gain in a single-stage R.F. amplifier: a reasonable amount of A.F. gain and a moderate power output are desirable, of course, if a loudspeaker is to be used.

A more detailed discussion of straight receivers is given in Chapter 4.

The Superhet Receiver

For high sensitivity, extreme selectivity and reliability, the superhet takes first place. The term " superhet," is a contraction of " superheterodyne," or to use the full expression, " supersonic heterodyne." The name is quite descriptive in that the fundamental principle of the receiver consists in the heterodyning of the incoming signals at a supersonic frequency; it is at this frequency that the main amplification and frequency selection take place.

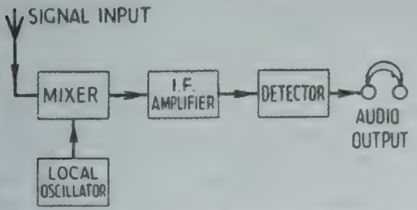


Fig. 16.
Block diagram showing the elements of a simple superhet.

Although superhet circuits appear to be somewhat complicated when compared with those of straight receivers, they can be resolved into relatively simple sections. Fig. 16 shows the elements of a superhet in the form of a block diagram. The first section is a frequency-changer, sometimes known as a *mixer* or *first detector*. The frequency-changer is supplied with two inputs, (i) the incoming signals and (ii) locally-generated R.F. oscillations. These two inputs are mixed together in the frequency-changer valve with the result that the local oscillator modulates the incoming signals, thereby producing in the output circuit of the frequency-changer a series of frequencies made up of the signal frequency and the oscillator frequency. Thus if F is the oscillator frequency and f is the signal frequency, the anode current contains components having the following frequencies:

$$(F+f) \quad (F-f) \quad (2F+f) \quad (2F-f) \quad (F+2f) \quad (F-2f) \; . \; . \; . \text{ etc.}$$

For various reasons, the most useful of these is $(F-f)$, that is to say, the *difference frequency* between the local oscillator and the signal frequency :

if f is greater than F, it is still the numerical difference that matters, and the choice as to which should be greater is determined by secondary considerations. The difference-frequency is filtered from all the others by a suitable tuned circuit in the anode circuit of the frequency-changer valve.

A simplified arrangement is shown in Fig. 17. Here V is the frequency-changer in which the first control grid G_1 injects the signal voltages from the signal circuit L_1C_1 : the local-oscillator voltage is applied to a second control grid G_3 (which is separated from G_1 by a screen grid G_2). The electron current

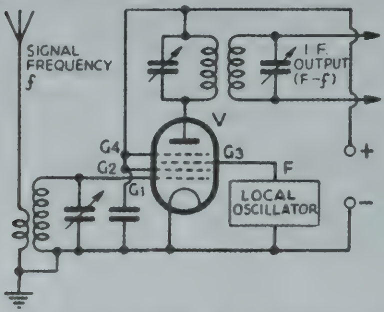


Fig. 17.
Simplified circuit arrangements for a hexode frequency-changer. The resonant transformer in the anode circuit is tuned to respond to the difference-frequency $F-f$

flowing to the anode is thus dependent on both the signal voltage and the local-oscillator voltage. In order to prevent coupling between the anode and the oscillator injection grid, a further screen grid G_4 is provided. The valve is thus a hexode. In the majority of receiver designs the local oscillator is combined with the frequency-changer valve, *e.g.* a triode-hexode. Another combined mixer-oscillator is the heptode. A more detailed discussion of frequency-changer valves is given in Chapter 5 and in *Valve Technique*, pages 56–63.

If the difference-frequency is made suitably low (*e.g.* between 50 kc/s. and 2,000 kc/s.), the output from the frequency-changer can be amplified to a very high degree and with a high measure of frequency selectivity. The amplifier which is used to boost the output from the frequency-changer is called the *intermediate-frequency amplifier* (usually abbreviated to *I.F. amplifier*). The term " intermediate " is used because, in this part of the receiver, the frequency is intermediate between the input signal frequency and the output audio frequency. One, two or three I.F. stages may be used, according to the required gain and selectivity.

Except in special superhets of elaborate design, the I.F. amplifier operates at a fixed frequency. This practice has many advantages, the most important of which is the very simple design. A typical I.F. amplifier stage is shown in Fig. 18. The four tuned circuits (two per transformer) require adjustment only when the receiver is being aligned, after which they may remain untouched.

The valves best suited to such amplifiers are high-gain R.F. pentodes or tetrodes, and if A.V.C. is to be used they should be of the variable-mu (remote cut-off) type.

A detector, or demodulator, is used to convert the modulated I.F. voltage into A.F. voltage. Any normal detector (see, for instance, Fig. 11) will serve this purpose. The output from it may be fed to an A.F. amplifier in the ordinary way.

It should be noted that in this form the detector will give audible signals only when the carrier is modulated. C.W. signals will not be audible. For the reception of C.W., it is necessary to inject an auxiliary oscillation of

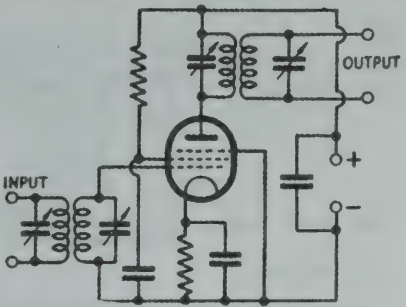


Fig. 18.
Typical single-stage I.F. amplifier. The transformers may be of the permeability-tuned type, in which case the adjustment is usually made by moving the iron cores and the shunt capacitances are fixed.

constant amplitude and frequency into the detector circuit so that it heterodynes the I.F. signals and thus produces an audible beat note. This auxiliary oscillator is known as the *beat-frequency oscillator* (B.F.O.). One way in which this can be coupled into the I.F.+detector combination is shown in Fig. 19.

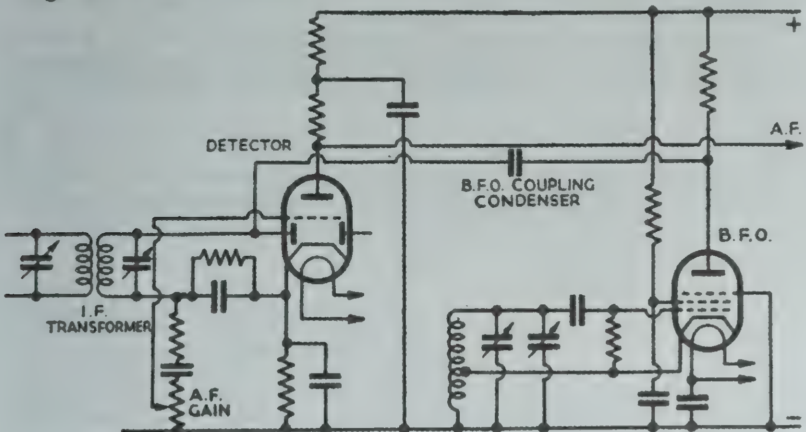


Fig. 19.
Typical second detector arrangement in a superhet showing how the B.F.O. is coupled to the detector input.

If the I.F. amplifier operates at, say, 450 kc/s., the B.F.O. should be adjustable over a frequency range of about 450–453 kc/s. (or alternatively 447–450 kc/s.), in order to provide a range of beat frequencies up to 3 kc/s.

For several reasons, which are explained in Chapter 5, it is highly desirable to introduce an R.F. amplifier in front of the frequency-changer valve. In many superhet designs two stages are used, and even three stages are sometimes considered worth while. For all moderate requirements one stage is sufficient—and far better than none at all.

A superhet comprising the several sections which have so far been mentioned is represented by a block diagram in Fig. 20.

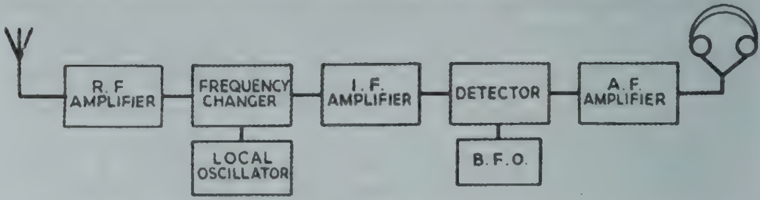


Fig. 20.
Block diagram showing the essential parts of a superhet provided with an R.F. amplifier and capable of receiving C.W. or telephony.

F.M. Receivers

The frequency-modulation receiver must be responsive to transient variation in frequency which occur in incoming F.M. signals, but should be insensitive to changes in amplitude. This is achieved by including a circuit specially designed to be frequency-selective (known as a *discriminator*), and by amplifying the signal until it is strong enough to saturate an amplitude-limiting device. In the F.M. receiver there is no need for a detector of the ordinary kind as found in A.M. receivers.

A block diagram of a typical F.M. receiver is shown in Fig. 21. It will be seen that the arrangement has much in common with a superhet designed for A.M. reception. In fact, from the aerial to the I.F. amplifier there need

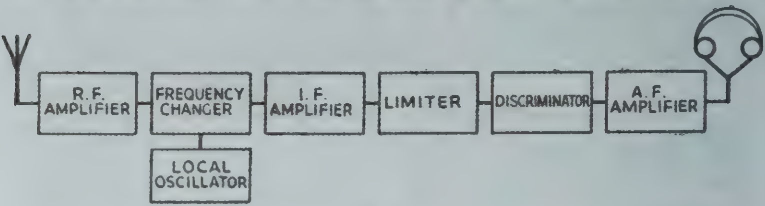


Fig. 21.
Block diagram of a typical F.M. receiver.

be no difference between the two types. The purpose of the limiter, which follows the I.F. amplifier, is to ensure that the signals passed on to the discriminator are free from amplitude modulation: this necessitates a rather higher amount of I.F. gain than is usual in A.M. superhets, and assumes that there will be no signals of smaller amplitude than that at which limiting occurs.

The limiter is very similar to a pentode I.F. amplifier stage except that it is operated with very low voltages on the screen and the anode. Under these conditions the valve is easily saturated by quite a small grid voltage swing. A more effective levelling of the signal amplitude is obtained by using two limiters in cascade.

The discriminator may take various forms, but the type in common use (known as the *Seeley* discriminator) has a centre-tapped tuned circuit and two diode rectifiers: see Fig. 22. The output from this circuit will have an

amplitude corresponding to the amount of deviation of the carrier frequency, and the modulation frequencies will be identical to the modulation frequencies in the transmitted signal.

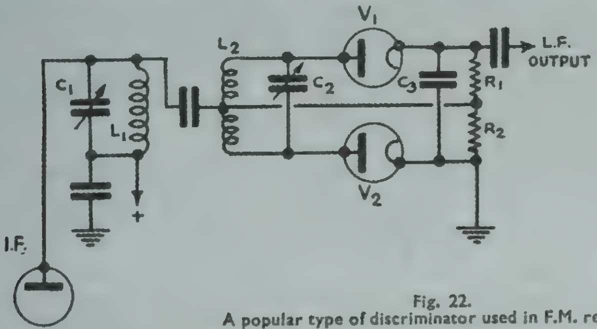


Fig. 22.
A popular type of discriminator used in F.M. receivers.

The A.F. amplifier which follows the discriminator is of conventional design.

Further details of F.M. receivers are given in Chapter 6 and in *VHF Technique*, pages 84–88.

CHAPTER 3 TUNING CIRCUITS

An inductance and a capacitance connected together possess the property of electrical resonance at a certain frequency depending on their respective values. The inductance and the capacitance may be connected either in parallel or in series, as shown in Fig. 23. The resonant frequency in either case is given by

$$f = \frac{10^6}{2\pi} \sqrt{LC}$$

where f is in kc/s., L is in microhenrys and C is in micromicrofarads. The frequency can be varied by changing either L or C or both. For several reasons it has been customary in the design of tuning circuits to keep L constant and to vary C where the tuning adjustment must be made with a high degree of precision. However, in some types of I.F. transformers and tuning circuits for V.H.F. receivers (above 40 Mc/s.), the adjustment is made by varying the inductance.

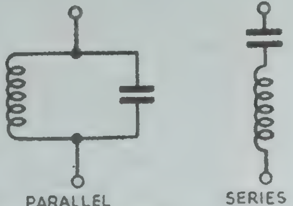


Fig. 23.
Elementary forms of resonant circuits

Variable Condensers

The equation giving f in terms of L and C shows that the frequency varies in proportion to the square root of L or C, so that a frequency range of 3 : 1, for example, requires an inductance variation—or a capacitance variation—of 9 : 1. In the conventional design of variable condenser, the variation is effected by a 180° rotation of the rotor vanes. The condenser plates can be designed

to give any desired rate of change of capacitance, *e.g.* straight-line capacitance, straight-line wavelength or straight-line frequency. In a straight-line capacitance condenser, the rotor plates have a semi-circular shape. It should be noted that the stator plates may have any shape, provided that they can afford a complete overlap of the rotor plates at full capacitance: see Fig. 24.

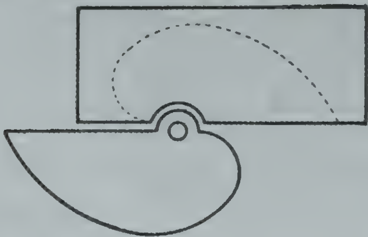


Fig. 24.
A variable condenser having vanes of this shape has a straight-line capacitance law.

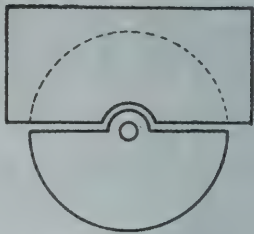


Fig. 25.
Vanes of this shape are used to provide a straight-line frequency variation.

This type gives a wavelength in variation of the square-law type if the stray capacitances in the tuning circuit are negligible compared with the capacitance of the variable condenser. This may hold at large-capacitance settings, but near the minimum the stray capacitances will be large enough to cause the frequency deviation to depart from the square law.

When a straight-line frequency variation is required, the capacitance law of the condenser should be a square law: see Fig. 25. Again the exact law of variation of frequency near the minimum setting will be dependent on the stray capacitances in the circuit.

Where the tuning circuits include padding and trimming condensers or band-setting condensers, the law of frequency variation will depart very considerably from that corresponding to the case when there is no capacitance other than that in the variable circuit. The design calculations for such circuits are often highly complicated.

Variable Inductances

During the last few years, variable inductances have come into widespread use in order to meet special requirements. The rotary type, in which the whole coil rotates on its axis against a stationary contact, is generally confined to transmitter circuits and is not well suited for use in receivers. There are two other forms, however, which offer real advantages in receiver design, *viz.*, the slug type and the magnetic-core type.

(i) *Slug-tuned Inductances.*—In this type a non-magnetic slug is placed in the field of the inductance and its position is varied along the axis by a screw thread: see Fig. 26A. The metal slug constitutes a short-circuited single turn, coupled to the coil, as represented in Fig. 26B. The effective inductance of the coil is thereby reduced, the amount depending on the tightness of the coupling. It is not easy to calculate the law of variation, and the design is usually based on experiment.

The losses are relatively small if the slug is of high conductivity: copper is best but brass is quite satisfactory.

The slug-tuned inductance is used largely in V.H.F. receivers where a wide-band characteristic is required, the associated capacitance being only that provided by the valves and the circuit wiring.

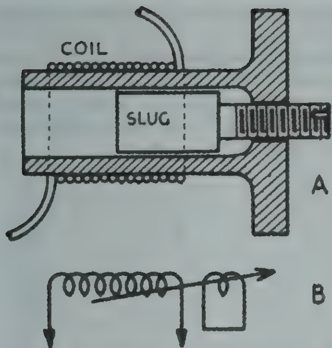


Fig. 26.
Slug-tuned variable inductance: (A) shows a cross-section of the solenoid and the non-magnetic core adjustable in position by means of the screwed shank: (B) shows how the slug is equivalent to a small short-circuited winding adjustably coupled to the solenoid. The sketch shown at (A) also represents the construction of a permeability-tuned (iron-cored) inductance.

(ii) *Iron-core Inductances.*—The construction of this type is almost identical with that of the slug type shown in Fig. 26, but there is one fundamental difference. Instead of a non-magnetic slug, there is a magnetic core made of compressed iron powder. The reason for using a powder is the avoidance of eddy-current losses which is achieved by breaking up the iron into individually insulated particles. Consequently there is no circulating current in the core and no tendency for the inductance to be reduced. On the contrary, the presence of iron in the field of the coil increases the magnetic flux and the inductance value is increased. As in the slug type, the law of variation is difficult to calculate, but unlike the slug type it has a positive characteristic; *i.e.* the inductance *increases* as the core is screwed into the field of the coil.

The magnetically-tuned coil is used largely in I.F. transformers and in wide-band V.H.F. tuning circuits.

Resonance

In a circuit containing inductance and capacitance, the amplitude and phase relationships between the voltages and currents depend on the frequency of the applied voltage. There are two simple but important cases: (*a*) the parallel circuit, and (*b*) the series circuit. These are shown in elementary form in Fig. 27.

In the parallel arrangement, at the resonant frequency the current is exactly in phase with the applied voltage. Assuming that the components are perfect, the impedance of the circuit at resonance is infinite: at frequencies above resonance the impedance has a finite value and the circuit has capacitive reactance, while below resonance the reactance is inductive.

In the series arrangement, at the resonant frequency the current which flows as a result of the applied voltage is a maximum: with perfect components the impedance is zero.

Effect of Resistance in Tuned Circuits

Any practical tuned circuit unavoidably contains resistance. If this resistance is increased the resonance becomes flattened, *i.e.* the resonance peak becomes less prominent, and if the resistance is relatively high it may be difficult to determine the precise adjustment for resonance because the peak has become so broad.

The effect of resistance is therefore to cause the circuit to discriminate less between a voltage at the resonance frequency and a voltage at a frequency

slightly different from it. In other words, the selectivity of the circuit is diminished as the resistance is increased.

The inductance and the capacitance both contribute something to the total resistance in the circuit: even if no resistance is deliberately included, there will be some resistance. If high selectivity is required, the coil and condenser should therefore be well designed. The size of wire, the winding dimensions

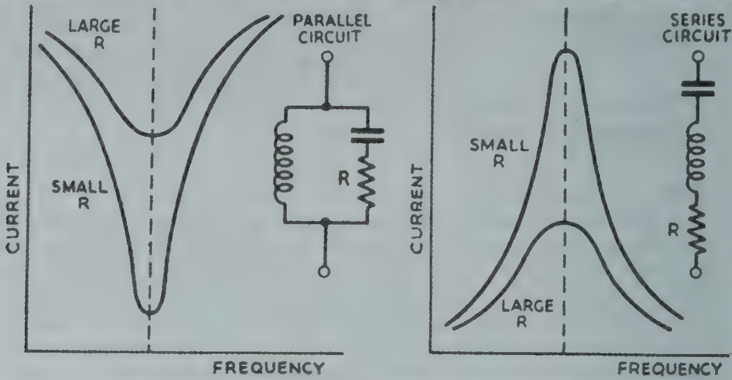


Fig. 27.
The resonance characteristics of parallel and series tuned circuits.

and the nature of the supporting material all affect the resistance of the coil. Condensers can generally be made with a very low effective resistance, and most of the losses in the normal type of tuned circuit are therefore in the coil.

Q (Quality Factor)

The "goodness" factor of a tuned circuit is known as Q (signifying *Quality*) and the value of Q enters into many calculations in the design of tuning circuits. The Q-value for a condenser is given by—

$$Q = 1/2\pi fCR$$

where f is the frequency in cycles per second (c/s.), C is the capacitance in farads and R is the effective resistance in ohms.

Since both C and R are relatively small, Q is very high. For most calculations, the value of Q for a condenser may be assumed to be infinity.

The Q-value for an inductance is given by—

$$Q = 2\pi fL/R$$

where f is the frequency in c/s., L is the inductance in henrys and R is the effective resistance in ohms.

It may be noted that $1/2\pi fC$ is the reactance of a condenser and $2\pi fL$ is the reactance of an inductance. It follows therefore that Q represents the ratio of reactance to resistance.

The resistance of the normal type of tuning coil cannot be made to have a value much less than 0·5 per cent. of the reactance, and there are many instances where the resistance is almost 10 per cent. of the reactance: *i.e.* the Q-value of a coil may be expected to lie between 200 and 10.

Any losses due to poor design of the coil will show up as a lowering of the selectivity. Such losses may be due to—

(*a*) poor dielectric material for the former,
(*b*) unsuitable wire size,
(*c*) unsuitable proportions of the coil (*i.e.* ratio of length to diameter or spacing of turns),
(*d*) moisture,
(*e*) resistance shunted across the coil (*e.g.* the input impedance of a valve),
(*f*) presence of metal near the coil (*e.g.* chassis or screening can),
(*g*) coupling to other circuits having appreciable losses.

In the design of television receivers, it is quite common to connect resistances across the tuning coils in order to obtain a broad resonance characteristic, but in communications receivers the emphasis is on the reduction of all forms of loss in the inductance design.

Wide-band circuits are sometimes used in pre-amplifiers in communications receivers and also in transmitters designed to operate over a wide band without necessitating adjustment of the tuning controls. The selectivity of a tuned circuit increases as the ratio of L/C decreases: on the other hand, the voltage developed across it is greater as the L/C ratio is made greater.

Effect of Screening

When a coil is surrounded by a screening-can there are two important changes: (i) the inductance is reduced, and (ii) the effective resistance is increased. Also the self-capacitance is increased, but this effect is not often appreciable. These changes are more pronounced if the screening-can fits more closely round the coil. The increase in resistance is greater (*i.e.* the Q becomes lower) if the screen is made of a metal of higher resistance. In practice the screen should be large enough to allow plenty of space round the coil. A construction as indicated in Fig. 28 is likely to be satisfactory for the

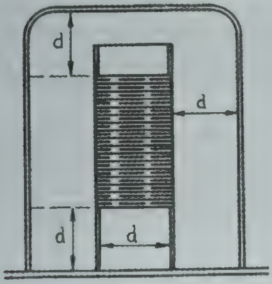


Fig. 28.
The clearance between a solenoid and its screening can should not be less than the diameter of the solenoid. The sketch shows the minimum dimensions for good performance.

most exacting conditions in high-selectivity circuits. The chassis and the screen may be of aluminium, brass or, better still, copper. Iron and steel are tolerable, but the effect of reducing the spacing will be more marked than if the higher conductivity metals were used, and larger clearances should therefore be allowed.

Impedance

If the capacitance associated with the inductance is relatively large (*i.e.* if the L/C ratio is small), the circuit will have a sharp resonance characteristic, assuming that the resistance losses are small, but if L is increased and C is decreased while the natural resonance frequency is kept the same, it is found that the selectivity is broadened. If the condenser is removed from the circuit, the inductance then offers an impedance over a very much wider frequency

range. It is, in fact, an untuned or aperiodic impedance, otherwise known as a *choke*. The coil will inevitably have a small self-capacitance, formed by the composite effect of the capacitances existing between the adjacent turns of the coil, and this means that the coil has a natural self-resonant frequency. Because the self-capacitance is very small, this resonant frequency is high, but it is not always so high that it can be disregarded. A large inductance when used as an aperiodic impedance (*i.e.* an R.F. choke) in a short-wave receiver may be found to resonate in the high-frequency tuning range. On the other hand, if the inductance of the coil is too small, it will not offer sufficient impedance over the required frequency range. Generally speaking, therefore, there is an upper and a lower limit of size for an inductance which is to serve as a choke coil.

The design of R.F. chokes is not a simple matter and it is better to work to an empirical rule: a useful rule is to wind one-third of a wavelength of fine wire in the form of a solenoid such that the ratio of length to diameter is at least 2: 1. The reason why it is desirable to have a high ratio of length to diameter in an R.F. choke coil is that the self-capacitance is thereby kept low.

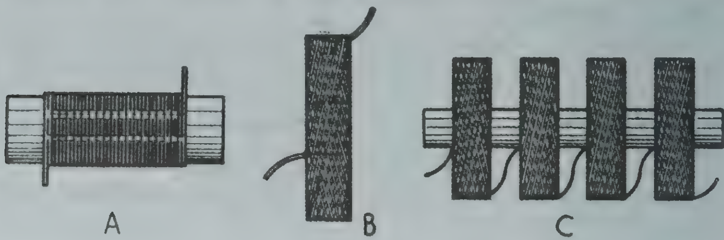


Fig. 29.
Typical R.F. chokes for use over various ranges of frequency. (A) Solenoid, e.g. 200 turns, ½ in. dia., 2 in. long. (B) Single pie-wound inductance. (C) Several pie-wound sections connected in series.

A slight spacing between adjacent turns also helps to keep down the self-capacitance. Multi-layer windings are permissible where a large value of inductance is required, but it is better then to split the inductance into series-connected inductively-coupled sections. Some typical R.F. chokes are shown in Fig. 29.

Methods of Calculating C, L and f

Capacitance.—The capacitance of a parallel-plate condenser is easily calculated. The formula is—

$$C = \frac{kA}{4 \cdot 45\, d}$$

where C = capacitance in micromicrofarads
k = dielectric constant ($k = 1$ for air)
A = total area of dielectric between the electrodes in square inches
d = distance between the electrodes in inches.

Example.—If $d = 0 \cdot 1$ in., and $A = 2$ sq. in., and the dielectric is air—

$$C = \frac{1 \times 2}{4 \cdot 45 \times 0 \cdot 1} = 4 \cdot 5\ \mu\mu\text{F}.$$

The formula gives the capacitance of a conventional type of variable condenser reasonably accurately, but it may be necessary to allow for the capacitance due to the metal frame. It is important, of course, to take the thickness of the plates into account when estimating the spacing between the electrodes.

The calculation of capacitance when the elements are irregularly shaped (*e.g.* valveholders, coil-selector switches) is hardly feasible. Nevertheless it is often necessary to have some idea what capacitances are contributed to a tuning circuit by the various circuit elements. There is such a wide variety of components that it is impossible to give accurate representative figures. The following typical values may be taken as an approximate guide:

Self-capacitance of tuning coil	5 μμF.
Coil-selector switch	3 μμF.
Input capacitance of valve	7 μμF.
Wiring	3 μμF.
Output capacitance of preceding valve	5 μμF.

This gives an equivalent capacitance across the tuning coil of 23 μμF. A complete calculation would have to include the minimum capacitance of the variable condenser: a typical value for the minimum capacitance of a normal 0·0005 μF. condenser is about 12 μμF., and to find this it may be necessary to add about 5 μμF. for the trimming capacitance.

Inductance.—The inductance of a solenoid is easily calculable. Several different formulæ have been developed, but the one due to H. A. Wheeler is probably the best suited to amateur needs. In its simplified form, this can be written as—

$$L = \frac{a^2N^2}{15\,(a + 3b)}$$

where L = inductance in microhenrys
a = coil diameter in inches
b = coil length in inches
N = total number of turns

Example.—If $a = \frac{1}{2}$ in., and $b = 1$ in., and $N = 20$, the inductance is—

$$L = \frac{(\frac{1}{2})^2 \times (20)^2}{15\,(\frac{1}{2} + 3)} = \frac{\frac{1}{4} \times 400}{15\,(7/2)} = 1{\cdot}9\ \mu\text{H}.$$

Frequency.—The resonant frequency of a tuned circuit is given by—

$$f = \frac{10^6}{2\pi\sqrt{LC}}$$

where f = frequency in kc/s.
L = inductance in μH.
C = capacitance in μμF.

Example.—If L = 15μH. and C = 35 μμF.,

$$f = \frac{10^6}{6{\cdot}28\,\sqrt{(15 \times 35)}} = 7350\ \text{kc/s}.$$

Wavelength.—The wavelength in free space is directly related to the frequency by the equation—

$$\lambda = V/f$$

where λ = wavelength in metres
V = velocity of electromagnetic waves in metres/sec.
f = frequency in cycles/sec.

The velocity in free space is approximately 3×10^8 metres/sec., or 186,000 miles/sec., but in the case of a wave travelling along the surface of a conductor, the velocity is somewhat smaller.

Example.—If $f = 7$ Mc/s. and if V is taken as 3×10^8 metres/sec.,

$$\lambda = \frac{3 \times 10^8}{7 \times 10^6} = 42{\cdot}86 \text{ metres}$$

Range of Frequency Variation

In medium-wave broadcast receivers, the tuning circuits are usually designed to cover the whole band, which extends from 550 kc/s. to 1,500 kc/s. This represents a ratio of approximately 3 : 1. Where a variable condenser is used to cover this range, the capacitance must vary over a range of $3^2 : 1$. In other words, the maximum capacitance must be nine times the minimum: this is about the highest ratio that can be achieved in a moderately-priced sturdily-built variable condenser.

In amateur receivers where the signal frequencies are higher and where the tuning bands are relatively narrow, the capacitance variation required is very much smaller. It is common practice to shunt the variable condenser with a fixed capacitance in order to reduce the frequency coverage to a desired small range.

One of the most important factors in receiver design is the *rate of change* of frequency in terms of the movement of the tuning control. If the frequency changes by much more than 1 kc/s. per degree of control-shaft rotation, the tuning adjustment will be found uncomfortably critical. Any tuning rate between 0·01 and 1·0 kc/s. per degree is satisfactory in amateur receivers.

In " all-wave " broadcast receivers, the variable condenser has to serve for covering a number of tuning bands of widely different frequency. This results in a rather tedious adjustment on the L.F. bands and a rather too critical adjustment on the H.F. bands. For easy handling and efficient operating, some means of " spreading " the frequency variation is a vital necessity.

Electrical Bandspread

Various modifications in receiver circuit design have been developed in order to overcome the disadvantage of cramped and congested tuning where a large frequency range is to be covered. One of these is the double-superhet. In this arrangement, the input circuits are allowed to be relatively broad in

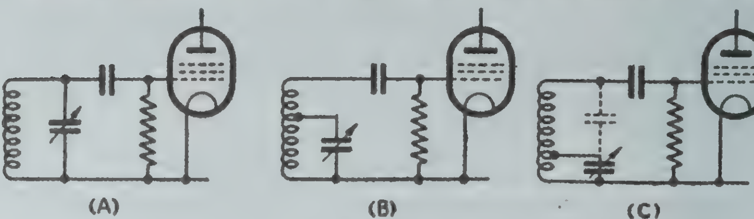


Fig. 30.

For wide frequency coverage, the tuning condenser is connected across the entire inductance as shown in (A). Bandspreading is achieved by connecting it across only a part of the inductance as shown in (B). If the tapping is made too low, as in (C), the stray capacitances shown dotted may cause spurious resonance effects.

heir tuning characteristic and the first local oscillator is adjusted to some suitable fixed value. The first I.F. amplifier (which has all its circuits gang-tuned) is tunable over the desired band. Of course, the frequency of the second local oscillator has to be varied at the same time so as to produce the necessary constant frequency-difference for supplying the input to the second I.F. amplifier.

A more simple method is the use of the tuning condenser across only a portion of the tuning coil: see Fig. 30. The calculation of the tapping point is not easy on account of the difficulty in measuring the stray capacitances, and the correct setting must be found by trial-and-error. If the tapping point is made very low, as indicated in Fig. 30C, there may be some spurious resonances in the tuning range, due to stray capacitances across the upper portion of the inductance.

Another method is the use of a larger fixed shunt capacitance across the variable condenser for the higher frequency ranges, whereby the effective maximum/minimum capacitance ratio is reduced: see Fig. 31. This has the disadvantage that on the higher frequency ranges, the effective capacitance across the tuning coil becomes unduly large in relation to the inductance, and the voltage applied to the grid may be quite seriously diminished.

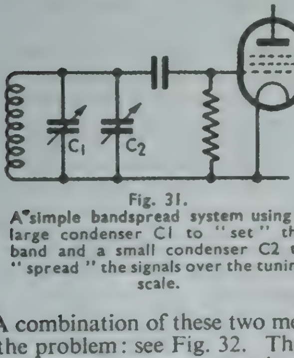


Fig. 31.
A simple bandspread system using a large condenser C1 to "set" the band and a small condenser C2 to "spread" the signals over the tuning scale.

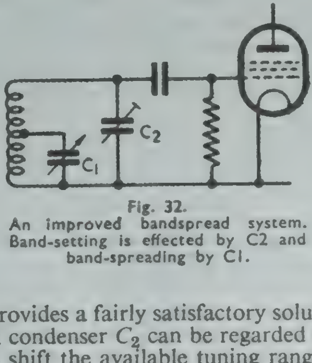


Fig. 32.
An improved bandspread system. Band-setting is effected by C2 and band-spreading by C1.

A combination of these two methods provides a fairly satisfactory solution to the problem: see Fig. 32. The pre-set condenser C_2 can be regarded as a *band-setting* adjustment: *i.e.* it serves to shift the available tuning range to the required portion of the spectrum. The variable condenser C_1 provides the tuning adjustment and may be called the *bandspread* condenser, *i.e.* it serves to "spread" the signals over a suitably wide range of adjustment.

Mechanical Bandspread

If a high-ratio reduction gear is used to couple the tuning condenser to the control knob, it is possible to achieve a satisfactory degree of bandspread without altering the circuit in any way. With this system, however, there is a temptation to use a large-capacitance condenser so as to cover a very wide range of frequencies with only one tuning coil, and at the low frequency end the sensitivity of the receiver will suffer on account of the large capacitance shunted across the inductance.

For good sensitivity, a capacitance of 100 $\mu\mu$F. is about the maximum that should be permitted across any tuning coil where the lowest frequency is about 1·5 Mc/s. At higher frequencies, the maximum permissible capacity should be correspondingly reduced.

The mechanical bandspread system throws into prominence the slightest imperfection in the mechanism. Any backlash in the condenser shaft will be evident, and likewise any backlash in the gear train.

Band-changing

Amateur communications receivers are usually designed to operate over a wide range of frequencies, *e.g.* 1·7–30 Mc/s., but this does not necessarily involve a continuous coverage. Some receivers are provided with complete coverage in order to permit the reception of short-wave broadcast transmissions, but for the amateur bands only, it is sufficient to provide relatively narrow bands, thus:

1·7–2·0; 3·5–4·0; 7·0–7·5; 14·0–14·4;
21·0–21·45; 28·0–30·0 Mc/s.

In either case—amateur bands only or complete coverage—it is necessary to include some means of changing the tuning coils. If one coil were used to tune from 1·5 to 30 Mc/s. (*i.e.* a frequency ratio of 20: 1), a capacitance ratio of 40: 1 would be required, and such a wide range would be extremely difficult to obtain in a single variable condenser. Moreover, the tuning range of 1·5–30 Mc/s. covers several thousand channels and the congestion on the tuning scale would be intolerable.

The receiver is therefore provided with means for changing the tuning coils according to the tuning range required. Three arrangements are possible: (i) interchangeable plug-in coils, (ii) switched coils, and (iii) rotating turret assembly.

(i) *Plug-in Coils.*—The plug-and-socket arrangement is the simplest system and is well suited to amateur construction. The coil former is integral with the plug, and the connecting leads can therefore be kept very short. Provided

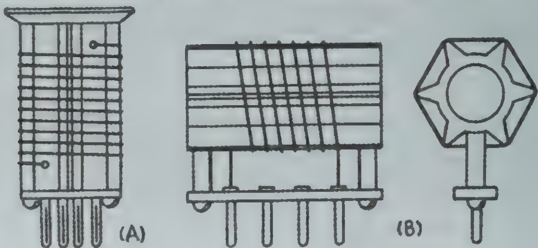


Fig. 33.
Popular types of plug-in coils.

good materials are used for the plugs and sockets, the efficiency attainable—as compared with the ideal arrangement of a built-in coil—is reasonably high. Fig. 33A shows the construction of a popular type of plug-in coil. For some circuits a 4-pin plug is sufficient, but 6-pin plugs are also in common use. Alternatively a transverse mounting can be used, with the pins placed in a straight line: see Fig. 33B.

Trimming condensers can be mounted individually on each coil to provide a means of accurately adjusting the tuning range. It is desirable to keep the condenser out of the field of the coil, and unfortunately the most attractive position—inside the coil—is the worst position electrically. Unless there are

serious constructional difficulties, a better position is outside the coil, with the main plane of the condenser plates running parallel to the axis of the coil: see Fig. 34.

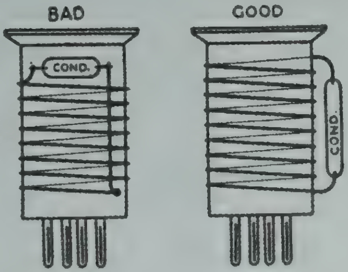


Fig. 34.
The best position for any condenser which is permanently connected to a coil is outside and parallel to the axis. If it is placed transversely inside the coil, the dielectric and eddy-current losses are at a maximum.

The former should be of the cylindrical ribbed type, the ribs serving to keep the wire as far as possible from the solid material in order to minimise the dielectric losses. Several good grades of material are available, and the choice becomes more important as the higher frequencies are approached.

Each plug-in coil may be fitted with its own individual screen, but it is much simpler (and equally effective) to place a fixed screen between the coil-socket positions, rigidly mounted on the chassis.

If desired, two or three coils may be fastened to a batten so that they can be plugged in or removed together. In principle this is a considerable advantage in band-changing, but it obviously requires close attention to mechanical details if it is to be entirely satisfactory in operation.

(ii) *Switched Coils.*—At one time it was generally considered that the losses incurred by the use of switches in the tuning circuits of a receiver were prohibitively heavy. In recent years, many highly successful receivers have been marketed having switched coils. If the switches are well constructed and if the groups of coils are arranged so that all the connecting leads are as short as possible, the efficiency attainable is practically as high as can be obtained by other methods.

The switches are preferably built with ceramic bases or wafers, and the metal parts should be suitably plated to ensure low contact-resistance. The paxolin-wafer type of switch can be used without serious loss of efficiency if ceramic switches of the desired kind are not available.

Careful attention must be given to the arrangements of the connections especially in regard to the short-circuiting of the coils not actually in use, This is important: if they are not short-circuited, there is a probability that at least one of them will resonate with the stray circuit-capacitances at a frequency within the selected tuning range, and by the coupling through the switch contacts or by inductive coupling may cause serious absorption effects. Fig. 35 shows an arrangement which usually proves satisfactory. All the coils which are of larger inductance than the one actually selected are short-circuited to earth by the rotating sector: the coils of less inductance than the one selected are not likely to resonate with any of the stray capacitances at a frequency within the range covered by the selected coil, and they may therefore be left free.

It is important to insert a screening panel between the switch wafers associated with the input and output circuits of an amplifier, in the same way as the tuning circuits are separated by a screen. A suitable arrangement of

switches, coils, screens and valves sockets for the R.F. amplifier, mixer and oscillator in a superhet is suggested in Fig. 36. This shows a screen between the signal-frequency circuit of the mixer and the oscillator circuit, an elaboration which is desirable in order to reduce " frequency-pulling."

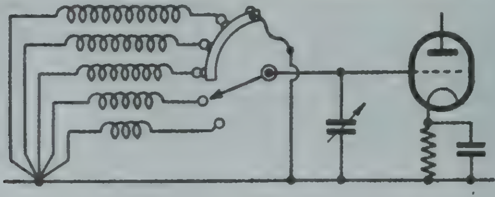


Fig. 35.
A coil-switching arrangement which avoids unwanted coupling effects by short-circuiting to earth all the coils larger than the one selected.

(iii) *Coil Turrets.*—In this system, the coils are mounted on a drum or barrel and each coil is brought into position against a fixed set of contacts. The constructional elements are shown in Fig. 37. As many as eight coils can be arranged in any one group without overcrowding. The important advantage is that long leads can be avoided: all coils have equally short leads.

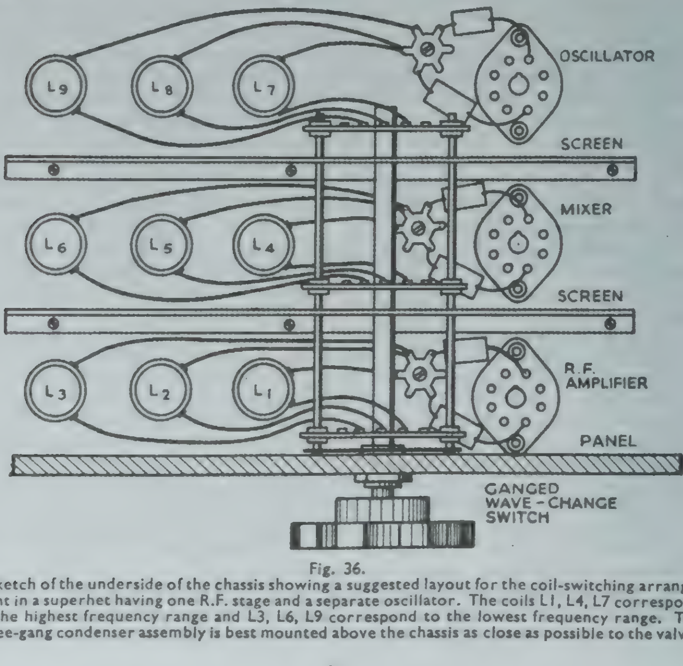


Fig. 36.
A sketch of the underside of the chassis showing a suggested layout for the coil-switching arrangement in a superhet having one R.F. stage and a separate oscillator. The coils L1, L4, L7 correspond to the highest frequency range and L3, L6, L9 correspond to the lowest frequency range. The three-gang condenser assembly is best mounted above the chassis as close as possible to the valves

Screening is, of course, necessary between the coils actually connected in circuit.

Electrically, the turret system is probably the best, but mechanically the ruggedness of the switches is offset by the rather large space required for the rotating coils. If it is necessary to experiment with different coils for one or more of the bands, the turret will provide easier access than the switched-coil system.

Ganged Tuning

In a straight receiver where there are two or more circuits to be tuned to the signal frequency, it is a relatively simple matter to link all the tuning adjustments together so that there is only one control. This is called *ganged tuning*. The most usual arrangement is to mount the necessary number of

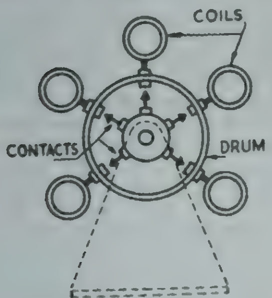


Fig. 37.
A turret mounting for the tuning coils in a band-switching receiver.

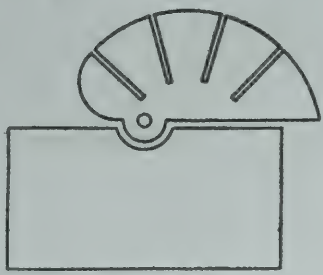


Fig. 38.
The end vanes of a tuning condenser are sometimes slotted. This allows the various parts of the vanes to be bent slightly by different amounts so as to modify the tuning characteristic.

variable-condenser elements on a common shaft. Individual adjustments to any one condenser section can be made by slightly bending the outermost vanes: radial slots are provided in most types for this purpose. Fig. 38 shows a typical split-vane condenser.

In a superhet receiver, the problem of reducing the tuning adjustments to a single control is complicated by the need for maintaining a constant frequency-difference between the signal-frequency circuit and the local oscillator circuit. If the receiver were required to operate on only one waveband, it would be quite feasible to design the oscillator section of the ganged-condenser to provide the frequency variation required to suit the signal-frequency circuits, and this has been done in some medium-wave broadcast receivers. Where two or more tuning bands are covered by the same condenser assembly, it is impossible to find a design which will suit the different requirements, and if a wide frequency-coverage is required it is generally considered necessary to resort to the use of padding and trimming condensers, as described in Chapter 5.

In amateur-bands receivers, the range of frequency on each band is so small (except perhaps on the 1·7 and 3·5 Mc/s. bands) that padding and trimming condensers are hardly necessary, and sufficiently accurate tracking can be achieved by choosing suitable values of inductance.

Impedance Transformers

The transfer of energy from one part of a circuit to another (*e.g.* from the aerial to the grid circuit of the first amplifier) depends on the relative impedances. Energy cannot be efficiently passed on from one part to the other unless the part which is supplying energy is properly loaded. In simple terms, a device which operates best under conditions of high voltage and low current (*i.e.* high impedance) is unable to absorb energy efficiently from a device which supplies the energy at low voltage and high current (*i.e.* low impedance). By inserting an impedance transformer between the two devices, the transfer of energy can be greatly improved. It is impossible to achieve the theoretical maximum efficiency owing to resistance losses and unavoidable radiation, but an efficiency of 60–90 per cent. is commonly obtained. Maximum efficiency occurs when the two impedances are equal.

Impedance transformers may be inductive or capacitive, or a mixture of the two. The inductive type is more frequently used. A simple example is the coupling of a dipole aerial to the grid circuit of an R.F. amplifier. If the energy in the aerial is taken from the centre point it will be in the low-voltage, high current form (*i.e.* low impedance). If taken from one end of the dipole it will be in the high-voltage low-current form (*i.e.* high impedance). For various reasons, the low impedance method is preferred.

The impedance value at the centre of a dipole is usually about 70 ohms, whereas the impedance of the input circuit of an R.F. amplifier may be about 10,000 ohms: see Fig. 39. To couple the valve efficiently to the aerial, an

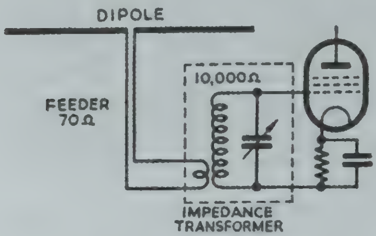


Fig. 39.
A transformer is necessary to transfer energy efficiently from a low-impedance feeder to the high-impedance input circuit of an amplifier. The simple type of transformer shown here is very popular and is quite effective.

impedance transformer is essential. It may also be regarded as a voltage step-up transformer (or a current step-down transformer). The essential requirement is to modify the impedance of 10,000 ohms so that it appears as an impedance of 70 ohms to the signals which are being brought in by the 70-ohm feeder.

A suitable transformer comprises a low-impedance primary (*e.g.* a 2-turn winding) and a high-impedance secondary (*e.g.* a 25-turn winding), with close coupling between them. The impedance ratio of the transformer is given by—

$$\left(\frac{\text{number of secondary turns}}{\text{number of primary turns}}\right)^2$$

In amateur practice the impedances are not often known with any useful degree of accuracy and it is generally necessary to find the optimum design by experiment.

The size of the secondary in Fig. 39 must also be considered in relation to the tuning capacitance. If a large tuning condenser is used, the inductance

will be relatively small and it will not be possible to obtain a very high turns ratio. Conversely, a high ratio is obtainable if the tuning capacitance is kept small.

Inter-valve R.F. couplings (*e.g.* I.F. transformers in superhets), are designed on the same principles, with additional emphasis placed on the degree of selectivity. To obtain high selectivity, the tuning capacitance must be reasonably large, and it may be preferable to sacrifice some of the voltage step-up obtainable with a high inductance secondary in order to maintain the desired sharpness of tuning.

Capacitive-type Impedance Transformers

The circuit shown in Fig. 40 illustrates the principle of the capacitive-type of impedance transformer. For any given frequency, the impedance of a

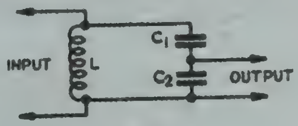


Fig. 40.
A capacitive type of impedance transformer. The circuit is reversible and the input and output connections may be interchanged, the impedance ratio being correspondingly reversed.

condenser is inversely proportional to its capacitance. Thus the impedance of the output circuit may be changed by using a different value for C_2: in order not to alter the resonant frequency of the circuit it will be necessary to use a different value for C_1, the value being chosen so that the effective capacitances of C_1 and C_2 in series remain the same. If C_2 is very small (high impedance) and C_1 is large (low impedance), the output circuit impedance will be nearly as high as the input circuit impedance, and if the values of C_2 and C_1 are interchanged, the output circuit impedance will be low in relation to the input circuit impedance.

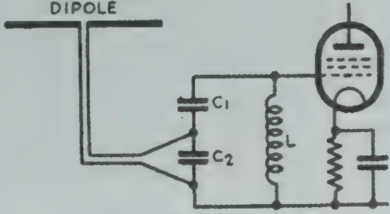


Fig. 41.
The capacitive type of impedance transformer represented in Fig. 40 is used here for coupling the low-impedance feeder to the high-impedance input circuit of the amplifier. A small R.F. choke may be connected across C2 to prevent spurious responses.

The terms *input* and *output* are used merely for convenience when describing the conversion of the energy as it passes through the associated circuits, but the theory is applicable in the reverse direction, and in this sense the two terms are interchangeable. This means that the circuit of Fig. 40 could be used for matching a low-impedance aerial feeder to any amplifier grid circuit by making C_2 suitably large and C_1 suitably small: see Fig. 41.

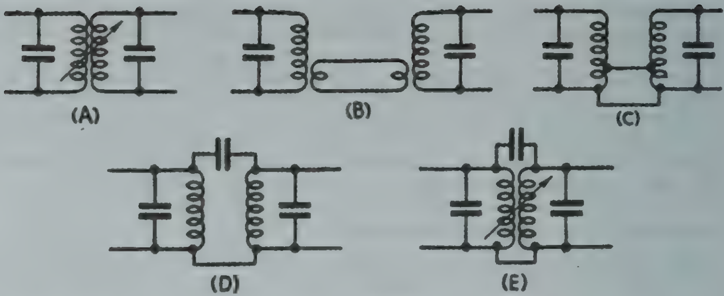


Fig. 42.
Various types of impedance transformers, using capacitive or inductive coupling or a combination of both.

The coupling between two resonant circuits may take various forms as shown in Fig. 42. Each has its own advantages and disadvantages, and the choice should be made after due consideration has been given to such details as—

(i) isolation of circuits (absence of D.C. connection),
(ii) range of impedance values,
(iii) tightness of coupling,
(iv) band-width characteristics,
(v) removal of harmonic frequencies.

CHAPTER 4 STRAIGHT RECEIVERS

IN spite of the numerous attractions of the superhet principle, the straight receiver is still popular, largely on account of its simplicity. It is cheap to build and easy to adjust for maximum performance, and there is no problem of image signals.

On the other hand, it is markedly inferior to the superhet in regard to selectivity (although it can be made reasonably selective on the 1·7 and 3·5 Mc/s. bands), and the sensitivity is somewhat lower. The straight receiver does not offer the same possibilities of accurate frequency calibration (unless it uses a separate oscillator instead of a regenerative detector), and the quality of speech reproduction is not very good (unless a relatively insensitive but distortion-free detector is used).

Selectivity

If a straight receiver is to have a high degree of selectivity, the design should include a sufficient number of tuned circuits resonating at the signal frequency, and each circuit should be operated under optimum load conditions and should have the minimum practicable L/C ratio. Since an R.F. amplifier usually has at least one tuned circuit associated with it, a sufficient number of tuned circuits can be achieved at least partly by introducing, say, two stages of R.F. amplification.

Following this line of reasoning, a layout such as that shown in Fig. 43 would be arrived at: this layout includes six tuned circuits. The selectivity

obtainable would be fairly good, even at frequencies as high as 7 Mc/s., but unfortunately serious difficulty would be experienced in the adjustment of so many tunable circuits if they were operated at maximum selectivity and

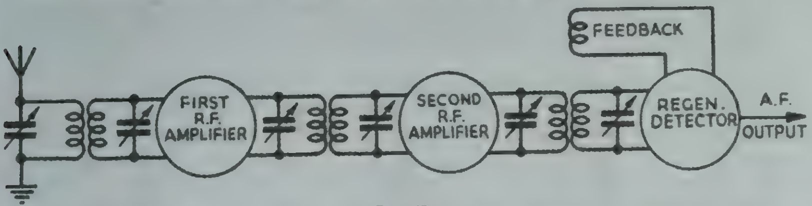


Fig. 43.
A theoretical form of receiver using two R.F. amplifiers with fully-tuned transformer coupling throughout. The six tuned circuits shown here would provide good selectivity but would be difficult to operate.

maximum gain. The simplified arrangement shown in Fig. 44 is more practicable. Here the number of tuned circuits has been reduced to three. The selectivity will be considerably reduced and there might be rather less gain overall, but the tuning control would be merely a 3-gang condenser and the alignment would be reasonably easy.

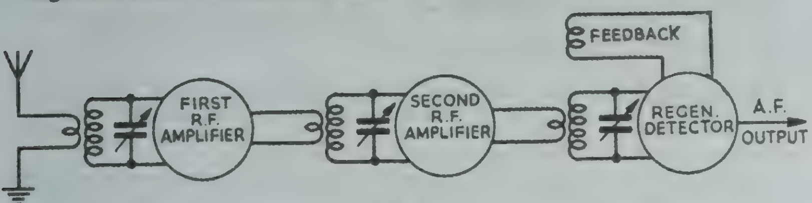


Fig. 44.
The arrangement of Fig. 43 simplified to a more practicable form. A three-gang tuning condenser can be used here if the inductances are carefully matched.

An equivalent arrangement using choke-capacitance coupling is shown in Fig. 45. This has the advantage of requiring fewer interchangeable connections for multi-band operation.

If the input impedance of the R.F. amplifier valves or of the detector is known to be relatively low, it may be worth while to provide a tapping on the tuning coils so as to avoid excessive loading of the tuned circuits: see Fig. 46. Such an arrangement is hardly likely to be necessary on frequencies below 20–30 Mc/s.

Much depends upon the choice of valve, but if modern R.F. pentodes (or tetrodes) are used, there should be no difficulty in obtaining satisfactory selectivity and gain. It should be remembered that the R.F. pentodes which are specially designed for television are intended for *broad band* operation (in which case a low input impedance is tolerable and even desirable), and therefore they may make it difficult to achieve the extremely high selectivity which is essential in a communications receiver.

The L/C ratio in the tuning circuits must be a compromise between maximum voltage step-up (high L/C ratio) and maximum selectivity (low L/C ratio), due consideration also being given to the frequency coverage required. Generally it will be found that if the tuning circuits are designed to

cover the specified frequency ranges with the smallest practicable size of variable condenser, the resultant conditions in the amplifier circuits will be sufficiently near the optimum.

The selectivity in any R.F. amplifier can be greatly enhanced by the use of positive feedback (*i.e.* regeneration). This can be provided by means of a small condenser (1–10 $\mu\mu$F.) connected between the control grid and the

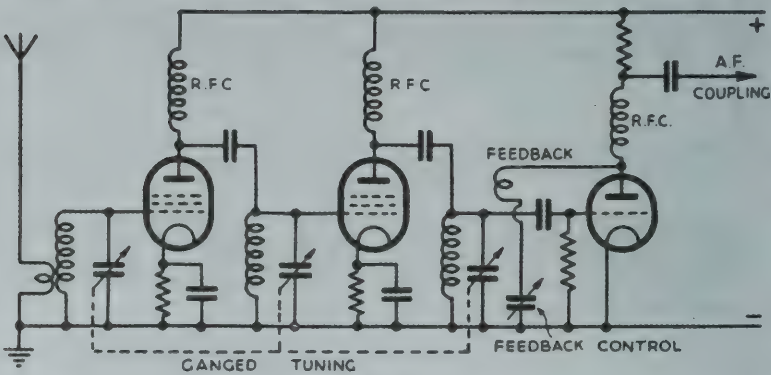


Fig. 45.
A modification of the circuit shown in Fig. 44. This has the advantage of fewer coil connections. A three-gang condenser may be used for single-knob tuning.

anode in the R.F. amplifier (thus partially cancelling the effect of the screen grid). Another method is to vary the by-passing of the feed to the screen grid: see Fig. 47. Here a variable resistance R is connected in series with the R.F. by-pass condenser C. When the resistance R is set at zero, the by-passing is completely effective and the feedback is at its minimum: when the resistance R is increased, the effectiveness of the by-pass capacitance is reduced and the feedback consequently increases.

A further alternative method is to use inductive feedback. The design and construction can be exactly the same as for a regenerative detector.

Any variation of the gain control affects the amount of feedback, and therefore the selectivity will depend not only on the feedback control but also on the gain control. Moreover, the amount of feedback obtained by any of these methods will almost certainly vary to a marked extent over the tuning range, and extra care will be necessary to get the best performance

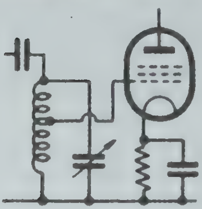


Fig. 46.
The effect of the input impedance of the valve on the tuning circuit can be reduced by connecting the grid to a tapping on the inductance. This improves the selectivity.

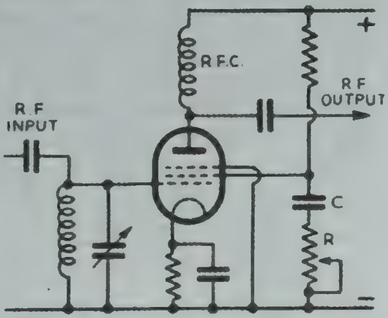
out of the receiver. In spite of these complications, the improvement in performance is generally considered to justify the extra difficulty in operation.

The condition known as single-signal reception is not possible in a straight receiver, since this requires a much higher order of selectivity.

Sensitivity

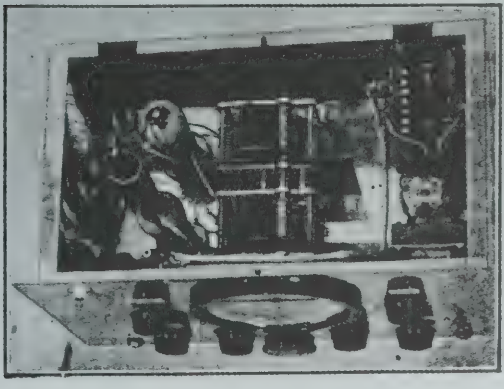
Even a simple regenerative detector receiving the input direct from the aerial without any R.F. amplification can be made remarkably sensitive. Its noise level is usually low and the signal/noise ratio can therefore be made

Fig. 47.
A method of controlling the gain in an R.F. (or I.F.) amplifier. The effectiveness of the screen grid in preventing R.F. feedback depends on the value of the resistance in series with the by-pass condenser. The optimum values for C and R depend to some extent on the frequency range in which the amplifier is operating and are best found by experiment.



quite high. Nevertheless, at least one stage of R.F. amplification is generally considered desirable for the reception of very weak signals. The gain in an R.F. stage diminishes as the frequency is increased, and unless care is exercised in the design and construction the gain may be found disappointingly low on the higher frequency bands. Even with two R.F. stages, it is hardly likely that there will be enough reserve gain to permit the use of A.V.C.

The sensitivity of a straight receiver is at its maximum for telephony reception when the regeneration in the detector is very nearly, but not quite, sufficient to cause self-oscillation. For C.W. reception, the sensitivity is at its maximum when the detector is only just oscillating; excessive feedback results in a loss of sensitivity due to the increase in the detector bias.



A 4-valve all-wave straight receiver for A.C. mains. Complete coverage from 120 kc/s. to 25 Mc/s. in seven ranges is effected by tandem connection of two sets of four-way coil switches. One R.F. stage is used with tuned-anode coupling.

Stability of Tuning

The most critical tuning adjustment in a straight receiver is the tuning of the detector circuit. This is affected by variation in the feedback adjustment, and the frequency calibration of the receiver will therefore be somewhat dependent on the regeneration. If the aerial input were fed directly into the detector grid circuit, the frequency would also be dependent on the degree of aerial coupling and would vary with movements of the aerial such as occur in a high wind.

An all-round improvement in stability can be achieved by the use of an R.F. amplifier stage. With careful design, the feedback conditions in the detector stage can then be made less dependent on the frequency adjustment, and the influence of the aerial can be completely removed.

If a very high order of frequency stability for C.W. reception is required, it will be necessary to use a separate oscillator: see Fig. 48. In this arrangement the C.W. heterodyne beat note is dependent entirely on the adjustment

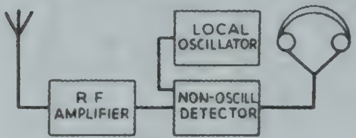


Fig. 48.
The use of a local oscillator for heterodyning C.W. signals in a straight receiver affords a high degree of frequency stability and permits accurate calibration.

of the local oscillator, and this can be calibrated with all the accuracy of a frequency meter. A moderate amount of regeneration in the detector is permissible (for either C.W. or telephony) but the detector must not be allowed to oscillate. The local oscillator should be of suitable high-stability type (see Fig. 49) and the strength of the oscillation fed into the detector should be carefully adjusted to give a reasonably strong beat note: excessive injection will reduce the sensitivity of the detector.

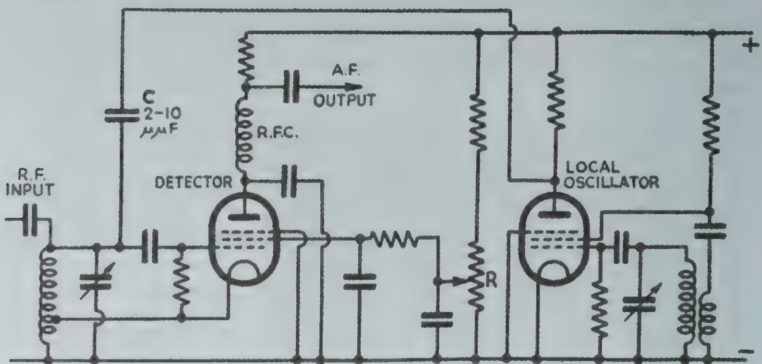


Fig. 49.
A typical circuit arrangement showing the coupling of the local oscillator (E.C.O. type) to the detector in a straight receiver. The strength of the injected oscillation can be adjusted by varying the coupling condenser C. A moderate amount of feedback is permissible in the detector but self-oscillation should not occur; the screen feed potentiometer R controls the amount of feedback.

A voltage-stabilised power supply is a great help towards achieving a high order of frequency stability. Such a supply need only feed the detector and and R.F. valves (or, for extreme economy, the detector and the R.F. screen grids only). This means that the voltage-stabilised supply need only provide 25–30 mA. at the most.

R.F. Amplification

At least one stage of R.F. amplification should be used. The best kind of valve for this purpose is undoubtedly a pentode (or the equivalent type of tetrode) and this may be either the variable-mu (remote cut-off) or short grid-base (sharp cut-off) type. Some of the more popular valves are—

1N5 6AK5 6J7 6SJ7 6SK7

EF50 EF54 KTW61 W81

Two forms of gain control suitable for R.F. amplifiers are shown in Fig. 50. Even if the overall sensitivity of the receiver is not very high, it is advisable

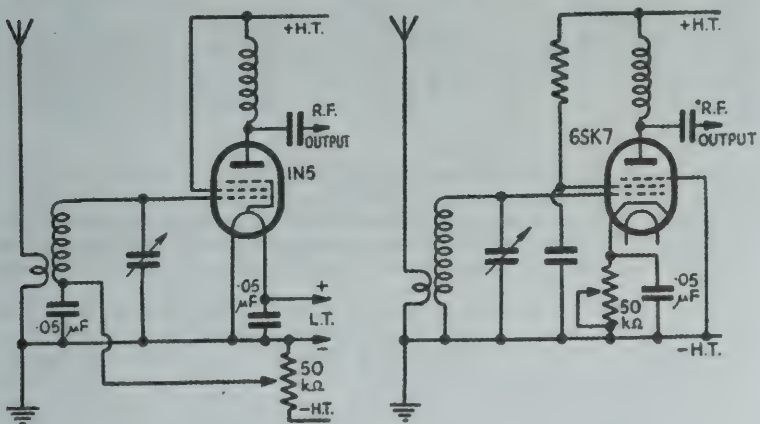


Fig. 50.
Two forms of gain control in R.F. amplifiers.

to provide a gain control in the first R.F. stage so as to prevent overloading by very strong signals.

It is usually necessary to enclose the valve in a screening can in order to avoid any stray pick-up. The filament or heater should be by-passed with a condenser of about 0·001–0·01 μF. All connections should be as short as possible and all by-pass condensers should be earthed at a common point to that particular stage.

If two or three stages are used, it is very desirable to make them identical in electrical characteristics and physical layout. This simplifies the adjustments for correctly tracking the multi-gang tuning condenser. Care is necessary in order to prevent unwanted feedback: adequate inter-stage screening must be provided.

It is not advisable to use more than three R.F. stages if they are all to operate under high-gain conditions. In television receivers, a wide frequency

band is essential and the stage-gain is unavoidably low: as many as five such stages may be used without fear of incurable instability.

The Detector

The overall performance of a straight receiver depends more upon the detector than upon anything else. Especially important is the control of regeneration at the threshold of oscillation. If there is any trace of *overlap* (sometimes called *backlash*), the maximum sensitivity will not be attained. This fault, which can be aptly described as " ploppy " feedback, is commonly met in beginners' receivers. It may be due to any one of a number of different causes or to a combination of them. The cure can usually be found by experimental variation of the circuit values, or in an obstinate case by changing to a different type of valve. The correct and incorrect threshold conditions are represented in Fig. 51.

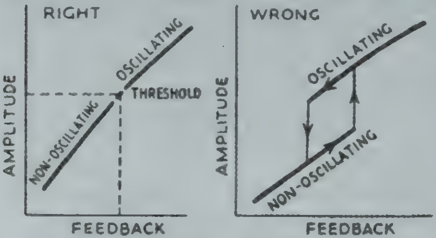


Fig. 51.
In a correctly adjusted regenerative detector circuit the threshold of self-oscillation is well defined and stable but free from abrupt changes.

The most important factors which are likely to need altering if overlap is present are the time-constant of the grid-leak/condenser combination, the anode voltage, the valve characteristics and the type of load in the anode circuit.

Although an R.F. pentode (or tetrode) offers the attraction of higher gain as a detector, a triode is usually found easier to control. For convenience in reducing the number of different valve types in use in the equipment, the detector is sometimes an R.F. pentode but with its screen (and possibly its suppressor grid) connected direct to the anode so that it functions exactly as a triode. A medium-gain triode is probably the best choice for satisfactory performance with the minimum of trouble, but even so it may be found that certain valves introduce an objectionable amount of mains-hum or are microphonic.

Some of the popular triodes which can be expected to function well as regenerative detectors are—

6C4 6J5G H63 MH4

The leaky-grid circuit is by far the best for a regenerative detector: see Fig. 52. The values of the grid condenser and the grid leak may have to be varied over the ranges of, say, 50–300 $\mu\mu$F. and $0{\cdot}25$–$3{\cdot}0$ MΩ in order to remove any trace of overlap.

The feedback coil L_2 (which is wound in the same direction as the signal coil L_1 and is coupled to the low-potential end of L_1) must not be much larger than is necessary to maintain oscillation. If L_2 is too large, overlap may be difficult to eliminate and the tuning of the grid circuit may be seriously affected. Generally L_2 has about one-third of the number of turns in L_1.

The regeneration control consists of the variable condenser C_2, sometimes

called the *throttle* condenser. The optimum size, which usually lies between 50 and 150 μμF., is best found by experiment. It is preferable to mount it well away from the panel, using an insulated shaft extension so as to avoid any hand-capacitance effects.

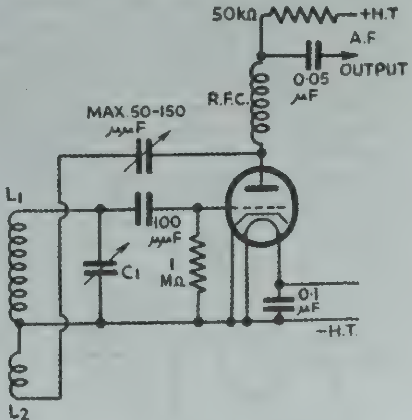


Fig. 52.
A typical leaky-grid regenerative detector circuit. Regeneration is controlled by the variable condenser connecting the feedback coil L2 to the anode. It is sometimes convenient to wind L1 and L2 as a single coil with one tapping point connected to the cathode. The number of turns in the feedback portion should be about one-fifth of the number of turns in the tuned grid portion. The values of the grid condenser and leak may have to be varied to remove the last traces of overlap.

An alternative method of regeneration control is the adjustment of the anode voltage by means of a potentiometer, but this is not to be recommended. If a tetrode or a pentode is used, it is quite practicable to control the regeneration by varying the screen voltage: see Fig. 53.

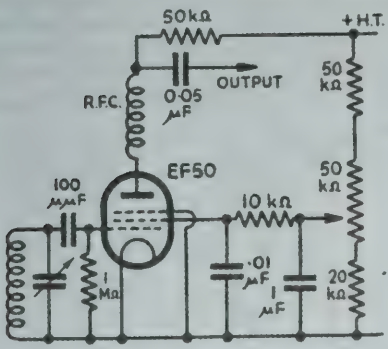


Fig. 53.
A method of regeneration or gain control in a detector or R.F. amplifier. The potential of the screen is controlled by means of a potentiometer connected to the H.T. supply, the supplementary fixed resistances being chosen so as to provide a convenient range of adjustment within the limits of the potentiometer. The 1μF condenser serves as a decoupling element and also smooths out irregularities in the action of the sliding contact. The 0·01 μF. condenser serves as an R.F. by-pass for the screen current.

The regeneration can also be controlled by varying the degree of coupling of the feedback coil. Although this method is capable of giving very smooth control of the threshold conditions, it obviously presents constructional difficulties where band-changing is required. It also tends to affect the frequency calibration rather seriously.

Any system of feedback control which relies on a potentiometer for the main adjustment is liable to give trouble due to wear and dirt on the sliding contact. If the potentiometer form of control is preferred (*e.g.* on account of the smaller effect on frequency) the potentiometer should be of the best

construction obtainable. A large by-pass capacitance is essential in order to smooth-out any minor irregularities that occur in the course of adjustment.

A.F. Amplification

For headphone reception it is hardly necessary to use any A.F. amplification. A single stage, comprising a small triode (*e.g.* H63, 6J5), will, however, provide a useful reserve of gain for boosting weak signals and for making reception more reliable in noisy surroundings.

It is preferable to isolate the headphones from the anode circuit for two reasons: (*a*) to prevent the flow of D.C. through the electromagnets, and (*b*) to avoid the risk of shock by contact with the H.T. supply. Resistance capacitance coupling is quite satisfactory: see Fig. 54.

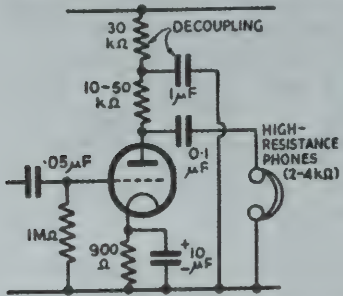
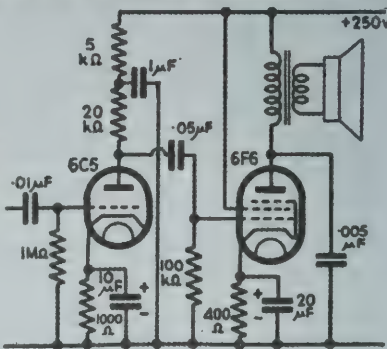


Fig. 54.
An output stage suitable for headphone reception. The triode may be a 6J5 or H63 or the equivalent. Transformer coupling to the grid may be used instead of the R—C coupling shown.

For loudspeaker reception the type of output stage obviously depends upon the acoustic power required. Unless the detector is specially designed to give low distortion (*e.g.* a diode rather than a leaky-grid triode) it is recommended that no attempt be made to produce a high-power output. A medium-size pentode or tetrode, such as a 6J7 or KT63, should be adequate for most purposes. A tetrode stage can be driven fully from the detector output,

Fig. 55.
An A.F. amplifier suitable for following a leaky-grid detector. Alternatively, the first A.F. stage could be a tetrode or a pentode, in which case the output stage could be a triode. Loudspeaker operation at high power levels is not recommended, on account of the distortion occurring in a leaky-grid detector.



although the signal may be boosted, if desired, by a single intermediate triode stage. Fig. 55 shows a suitable circuit giving the maximum gain that can be used satisfactorily after a leaky-grid detector.

Further details of A.F. amplifiers are given in *Valve Technique*, pages 23–30.

CHAPTER 5 SUPERHETERODYNE RECEIVERS

THE superheterodyne—abbreviated to superhet—is undoubtedly the most popular type of receiver for amateur use. It is inherently more complicated in design than the straight receiver. These two factors together provide the chief reason why most amateurs nowadays prefer to buy their receivers. At one time it was extremely rare to find a commercially manufactured receiver in an amateur station: it was even considered to indicate a lack of genuine interest, and a home-built receiver was usually credited with giving better performance.

There are some commercially produced receivers which are nowadays capable of giving superb performance which it would be difficult to excel. Nevertheless, it is quite possible for an amateur to design, construct and adjust a superhet embodying all the most modern principles which would be equal in performance to its commercial counterpart and might even be superior to many receivers of the lower-priced class. Refinements in the construction and adjustment of a receiver from the manufacturer's standpoint are very costly, while the amateur constructor can include them with relatively little effort or outlay.

Frequency Changer

It is generally appreciated that the most important part of a superhet is the *frequency changer* (sometimes known as the *mixer* or *first detector*). The valves commonly used (*e.g.* 6K8, X61, 1R5) are either triode-hexodes or heptodes and are tolerably effective. Typical circuit arrangements for such valves are shown in Fig. 56.

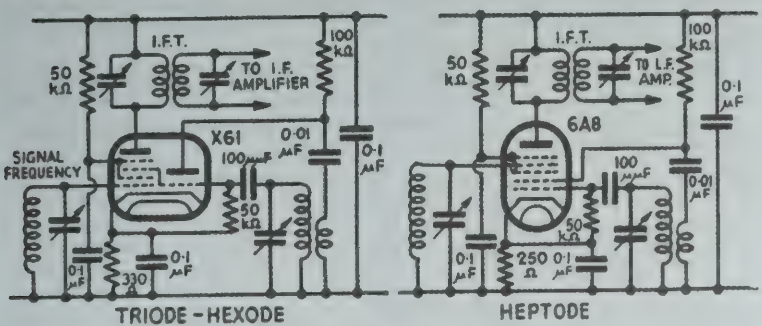


Fig. 56.

Typical frequency-changer circuits. Other arrangements of the valve electrodes are possible (as for example in the popular 6K8) but the circuit connections are similar in principle.

These circuits are shown with inductively-coupled tuned-grid oscillator circuits, but several other feedback arrangements may equally well be used. A Colpitts oscillator circuit is sometimes recommended: see Fig. 57. This is especially suitable for V.H.F. work on account of its rather higher stability. A refinement of this, known as the Clapp oscillator (Fig. 58), offers even greater stability for frequencies up to about 7 Mc/s.

Multi-electrode mixer valves tend to generate a troublesome amount of noise, due to the mechanism of the electron flow through the various grid electrodes: see *Valve Technique*, pages 82–86. Further, the miniature

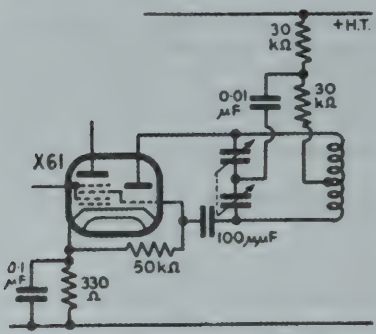


Fig. 57.
A Colpitts oscillator is sometimes preferred for a frequency-changer, especially for V.H.F. operation. The decoupling condenser and the rotor of the twin-gang tuning condenser should be shown connected to earth.

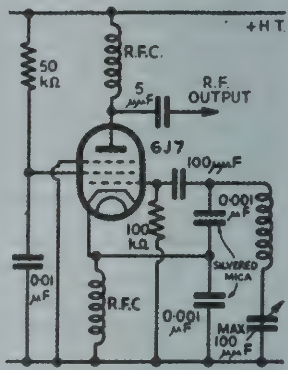


Fig. 58.
A Clapp oscillator, designed to have specially high frequency-stability. It tends to lose its high stability at frequencies above about 4 Mc/s.

electrodes in the oscillator section are barely large enough to provide an entirely satisfactory performance, *i.e.* without frequency drift and with constant voltage output over a wide tuning range. On account of this it is preferable to use a separate oscillator valve of reasonable proportions and to limit the function of the " frequency-changer " to that of mixing the incoming frequency with the locally generated frequency.

A valve which introduces less noise as a mixer than the multi-electrode types is the triode. No question of neutralisation arises because the anode and the grid circuit resonate at quite different frequencies and there is consequently no tendency towards self-oscillation.

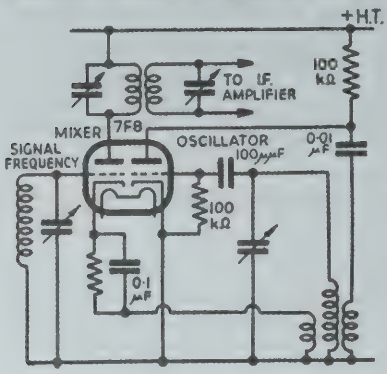
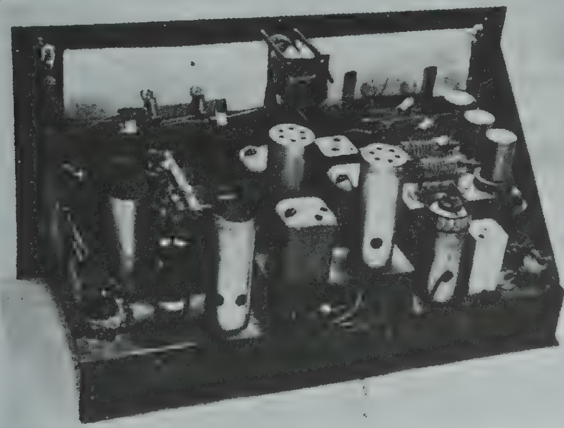


Fig. 59.
A double-triode frequency-changer. The noise generated in this arrangement, using a triode for mixing, is somewhat less than that in heptodes or triode-hexodes.

There are available on the market double-triode valves (*e.g.* 6C8G, 7F8) which can be arranged to constitute a highly satisfactory mixer-oscillator. A suitable circuit is shown in Fig. 59.

The question of frequency stability cannot be over-emphasised. The frequency should be independent of temperature, the supply voltage and the valve characteristics, and the frequency adjustment should, of course, not be affected by vibration and jolting.

Mechanical stability can be achieved by suitably rigid construction and by a careful choice of components (*e.g.* a variable condenser with thick vanes, widely spaced, preferably of aluminium). The questions of supply voltage and temperature are best settled by proper attention to the circuit design.



An amateur-built double superhet using switched coils for 1·7—14 Mc/s. and plug-in coils for the higher frequency bands. The 465 kc/s. I.F. amplifier is located centrally on a raised sub-chassis.

The coils should be wound tightly on low-loss formers, preferably of ceramic material with a dielectric constant having a low temperature coefficient. The metal screens used to prevent unwanted capacitance coupling and pick-up should be spaced as far as possible from the field of the coils.

In amateur-bands receivers, the frequency coverage on each band is relatively small, but where the coverage is large, special care may be necessary to ensure that the oscillator output voltage is kept as nearly constant as possible over the tuning range. This may be examined by means of a valve voltmeter or a cathode-ray oscillograph. If the voltage is found to vary appreciably, it may be stabilised by inserting a resistance of, say, 1,000 ohms in series with the feedback coil and perhaps also by changing the size of the coil. These factors cannot conveniently be prescribed, because they depend on the type of valve, the operating voltages, the design of the coils and other elements in the circuit.

The *pulling effect* (*i.e.* the effect on the oscillator frequency of the tuning of the signal-frequency circuit) is not likely to be appreciable if the I.F. is high enough to make the signal frequency sufficiently different from the oscillator frequency, or where the mixer circuit is of such a kind that there is negligible electrostatic or electronic coupling between the circuits. With an I.F. of the

order of 400–500 kc/s., there is little likelihood of serious pulling except perhaps when the signal frequency is higher than 15–20 Mc/s.; but for such frequencies it would be desirable to use a higher frequency in the I.F. amplifier (say 5 Mc/s.) with the object of avoiding images, and the trouble is then not likely to occur.

An ideal local oscillator should not generate harmonics. If harmonics *do* exist, they will be heard as continuous carriers in one or other of the tuning bands. Keeping the oscillator voltage as low as possible consistent with good efficiency and using thorough screening of the oscillator helps to prevent the harmonics from causing serious whistles. For V.H.F. work, it is sometimes advantageous to use the second harmonic of the oscillator to mix with the incoming signal because the fundamental frequency of the oscillator is thus kept relatively low and there is consequently less difficulty in minimising frequency drift.

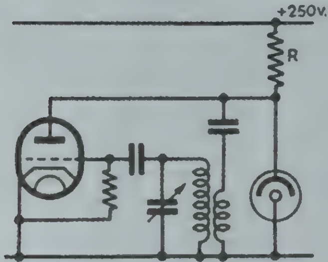


Fig. 60.
A simple voltage-stabiliser providing constant voltage for the local oscillator in a superhet. Suitable regulator tubes are VR75, VR105 (American), S.130 (*Cossor*), 7475 (*Mullard*). The resistance R must be adjusted to suit the operating conditions. Another resistance of about 50,000 ohms (not shown) should be inserted in the lead from the stabiliser tube to the anode circuit.

In many oscillator circuits, the frequency will be found to vary quite appreciably with the supply voltage, and it is desirable to include a voltage stabiliser for the oscillator supply. A suitable circuit is shown in Fig. 60.

Temperature Compensation

To overcome the drift of the oscillator due to the warming up of the circuit elements after the receiver is switched on, some designers have resorted to temperature compensation. In any receiver, each component associated with the tuning circuits has a certain temperature coefficient of frequency variation, depending on the thermal expansion of the various elements (coil windings, formers, spacers and insulators in variable condensers, valve holders, etc.). The composite effect of the several temperature coefficients results in a frequency drift, usually in a downward direction. By rearranging the circuit so that part of the tuning capacitance is provided by a special condenser having an opposite temperature coefficient, the overall effect of a change in temperature can be virtually eliminated. These special condensers are made with a carefully selected dielectric material.

The degree of compensation obtainable depends on numerous factors and must be considered as a separate problem on each frequency band. Very careful measurements must be made of the frequency drift that is to be counteracted in order to decide what fraction of the capacitance should be provided by the compensating condenser.

Tracking

If separate tuning controls are used for the oscillator and the signal-frequency (S.F.) tuning circuits, the question of tracking does not arise, for

the circuits can always be independently adjusted as required. Usually, however, the tuning of a superhet is controlled by a single knob and the oscillator and signal-frequency tuning circuits have to be designed so that they have a constant frequency-difference irrespective of the tuning control position or of the tuning band. The maintenance of this constant frequency-difference is called *tracking*.

The problem becomes more difficult as the ratio of the I.F. to the S.F. becomes higher, and as the ratio of the maximum to the minimum S.F. (*i.e.* the proportional tuning range) becomes higher. In a conventional broadcast receiver operating on medium frequencies, very substantial measures have to be taken to ensure correct tracking. In such receivers, the ratio of the I.F. to the S.F. is high (ranging from 0·3 : 1 to 0·9 : 1) and the ratio of the maximum to the minimum S.F. is high (approximately 3 : 1). On the other hand, in a receiver operating on the 3·5 Mc/s. band with an I.F. of 465 kc/s., the ratio of the I.F. to the S.F. is of the order of 0·465/3·5 (*i.e.* 0·13 : 1) and the ratio of the maximum to the minimum S.F. is 4/3·5 (*i.e.* 1·2: 1). Both of these ratios are low enough to make the problem of tracking relatively simple. For a 14 Mc/s. receiver with an I.F. of 465 kc/s. and an S.F. range of 14·0–14·4 Mc/s., the problem is so slight that it can be ignored. But if the 14 Mc/s. receiver has an I.F. of, say, 5 Mc/s. (*i.e.* a ratio of I.F. to S.F. of 1 : 3), it may be worth while to give the circuit design rather careful thought.

The loss of signal strength due to incorrect tracking depends on the selectivity of the signal-frequency circuits. Clearly, if the S.F. circuits have a broad resonance characteristic, the effect of mistuning from the resonance peak will be less marked than if the circuits are sharply resonant, but this should not be taken as an argument for low selectivity in the S.F. circuits. High selectivity is still desirable in order to reduce interference and cross-modulation.

Because the selectivity of the S.F. circuits in various receivers may range within quite wide limits, it is not possible to state in general terms what degree of error in tracking may or may not be permitted.

Tracking Circuits

In many receivers there are two or even three circuits tuned to the signal frequency while there is only one tuned to the oscillator frequency. For reasons of economy it is preferred to insert the tracking elements into the oscillator circuit. The basic circuit arrangement as used in a typical broadcast receiver is shown in Fig. 61. The oscillator frequency is set on the high

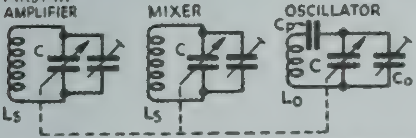


Fig. 61.
The basic tuning circuits of a superhet having one R.F. stage and a ganged frequency control.

side of the signal frequency but must change by only the same amount between maximum and minimum as the signal-frequency circuits. This means that the *proportional* coverage of the oscillator circuit must be reduced—assuming that the variable condenser is of the same type and size as those in the S.F. circuits. The effective capacitance range of the variable condenser C is reduced by the series condenser C_p (known as the *padding condenser*). This

has a more pronounced effect at the maximum capacitance. The effective capacitance at the minimum setting is suitably adjusted by means of the small parallel trimming condenser C_0. To achieve the required tuning characteristic, the oscillator inductance L_o must be reduced relatively to the S.F. inductance L_s.

The tuning curves obtained with a circuit arrangement of this form are shown in Fig. 62. The actual oscillator frequency curve does not lie exactly

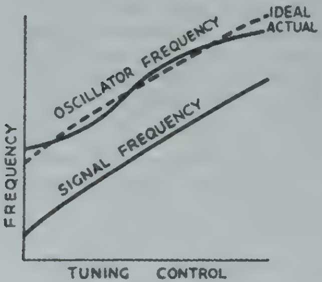


Fig. 62.
The circuits shown in Fig. 61, when correctly adjusted give a tuning characteristic of the form shown here. At three settings of the control the circuits are correctly tracked, but at all other points a slight error is inevitable. In this drawing the deviations are exaggerated for clarity.

at a constant frequency-difference above the S.F. curve, but crosses the ideal oscillator curve at three points. The positions of these three points and the amount of deviation from the ideal curve are determined by the settings of the padding condenser and the trimming condenser and the value of the oscillator inductance.

In amateur-bands receivers, it is quite unnecessary to provide a padding condenser. Calculations show that it should be extremely large—so large, in fact, that it can safely be replaced by a direct connection. The oscillator inductance should, however, be slightly smaller than the S.F. inductance.

Several different formulæ and charts have been devised to facilitate the design of tracking circuits, but in all of them it has been necessary to make so many assumptions that the calculated component values invariably have to be adjusted by trial-and-error to achieve a satisfactory performance. References are given in the *Selected Bibliography* for those who wish to study the analytical methods, but practical suggestions which it is hoped will be of more direct use are given in Chapter 7.

The detailed procedure for the alignment of the tracking circuits is described in Chapter 9.

I.F. Amplification

The I.F. amplifier has two main functions, (*a*) to amplify the incoming signals, and (*b*) to separate the wanted signals from the unwanted signals. It must therefore have high gain and high selectivity. In many receivers the gain and selectivity are not adjustable by panel controls, but it is distinctly advantageous to be able to vary both. The only reasons for not providing these controls are the reduction of cost and the saving of panel space.

Both the degree of amplification and the selectivity are dependent on the resonant frequency of the I.F. amplifier, and the choice of frequency must be carefully considered in the general planning of the receiver. If the I.F. is very low, *e.g.* 50 kc/s., the degree of amplification obtainable is very high

and the selectivity can be made extremely narrow. Unfortunately, second-channel (*i.e.* image) interference is likely to occur, since the image frequency will be very close to the signal frequency. Tracking difficulties will be negligible.

On the other hand, if the I.F. is very high, *e.g.* 10 Mc/s., it will not be so easy to obtain a high degree of amplification and the selectivity will be poor. Second-channel interference will be greatly reduced, but the problem of tracking will require careful attention.

These comparative characteristics are set out in Table I.

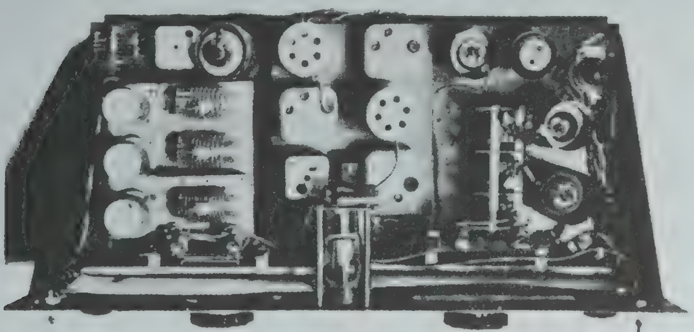
TABLE I

I.F.	Amplification	Selectivity	Image interference	Tracking
Very low (*e.g.* 50 kc/s.)	Good	Very high	Bad	Easy
Medium (*e.g.* 450 kc/s.)	Medium	Medium	Medium	Easy
Very high (*e.g.* 10 Mc/s.)	Poor	Poor	Negligible	Fair

For general purposes a frequency of about 450 kc/s. is a satisfactory compromise. Most manufacturers seem to prefer 465 kc/s., while 560 kc/s. has been used successfully in Service receivers. The exact value is of no great importance, but, of course, once the value has been decided upon, the several windings of the transformers must be adjusted carefully to suit this frequency (either " on the nose " or staggered so as to produce a band-pass effect).

Another value of I.F. which is sometimes used in commercial receivers is 1,600 kc/s. This affords good second-channel rejection (since the image is then separated from the true frequency by 3·2 Mc/s.) but the selectivity in the I.F. amplifier itself (*i.e.* adjacent-channel selectivity) is not very high. Moreover it is desirable to provide three stages of I.F. amplification if frequencies as high as this are used: the performance would then be comparable with that of a two-stage amplifier operating at, say, 465 kc/s.

The frequency chosen for the I.F. should be one that is not used by high-power stations: in this way the risk of direct pick-up of unwanted signals by the I.F. amplifier can be avoided.



A top view of the superhet illustrated on p. 45.

Double Superhets

By dividing the I.F. amplifier into two sections, one operating at, say, 10 Mc/s. and the other at 50 kc/s.—with suitable frequency changing—it is possible to secure the advantages of both and to avoid their drawbacks. This double frequency conversion is particularly advantageous for signal frequencies above about 15 Mc/s. The first I.F. should be at least 3 Mc/s. (and may be as high as 10 Mc/s.), and the second I.F. may be as low as 50 kc/s. (although frequencies of the order of 450 kc/s. are satisfactory).

Thus it becomes practicable to use a standard receiver as the first and second I.F. amplifiers: see Fig. 63. All that is necessary is to convert the incoming signals (by means of a frequency-changer and local oscillator) to

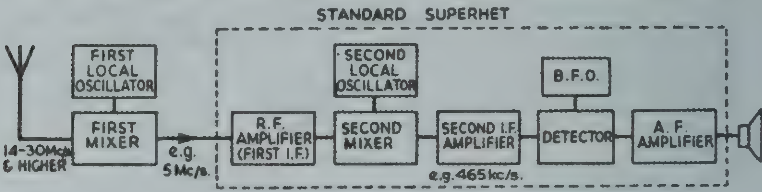


Fig. 63.

Block diagram of a double superhet. The signals are first converted to a selected frequency, such as 5 Mc/s., and are then fed into any conventional type of superhet, the input of which is tuned to this frequency. The I.F. in the main part of the receiver may be 100 kc/s. or less in order to obtain extremely high adjacent-channel selectivity.

whatever frequency is chosen for the standard receiver (*e.g.* 5 Mc/s.). There is a risk of interference between the first oscillator and the oscillator in the standard receiver, and adequate screening must be provided to ensure that the harmonics of the latter do not reach the first frequency-changer circuit. Such interference would be apparent as continuous, unmodulated carriers.

A suitable coupling arrangement for feeding signals into the standard receiver is shown in Fig. 64.

If the selectivity of the signal-frequency circuits is not too high, it is possible to use the tuning of the standard receiver as a band-spread adjustment, the band-setting being provided by the first oscillator and signal-frequency circuits. The anode circuit of the first frequency-changer (shown pre-set at

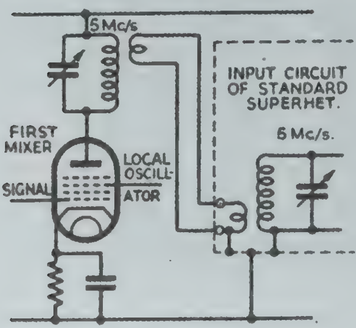


Fig. 64.

A low-impedance link coupling may be used to couple the output from the first mixer to the input terminals of a conventional type of superhet.

5 Mc/s. in Fig. 64) must then be variably tuned to match the adjustment of the standard receiver or must be given a suitably broad resonance characteristic by using a high L/C ratio and perhaps also shunting with a resistance (*e.g.* 5,000 ohms).

Variable Selectivity

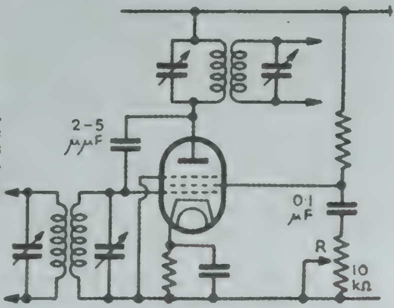
It is the selectivity of the I.F. amplifier that provides adjacent-channel selectivity. In C.W. receivers this is usually made as high as possible; so high, for example, that signals having a frequency separated by only 0·3 kc/s. from the wanted signal will be greatly attenuated. For telephony, severe distortion would result from such high selectivity owing to the serious loss of the upper audio frequencies. Even for C.W. reception a band-width of 0·5 kc/s. is often inconvenient, *e.g.* when searching or when the signal is not absolutely steady. It is therefore very desirable to introduce some form of adjustment of the selectivity in the I.F. amplifier, and most good receivers include this feature.

The simplest means of providing variable selectivity is a variable resistance connected across one of the transformer windings. This is not a convenient method on account of the awkward design problems in avoiding unwanted capacitance effects and in providing adequate screening of the leads to the potentiometer. It is also rather objectionable because of the loss of gain that results from the use of a resistance. Where a fixed degree of broader selectivity is required, a fixed resistance can be used and some of the drawbacks disappear: this method has been adopted in some television receivers.

Regeneration Control.—A better method of providing variable selectivity is the use of *controllable regeneration*. The circuit shown in Fig. 65 illustrates

Fig. 65.

Controlled regeneration in an I.F. amplifier. The variable resistance R determines the degree to which the screen potential is stabilised at R.F. and thus controls the feedback. A small condenser between grid and anode assists in providing slight positive feedback.



how this control can be made by means of a potentiometer R. This method has the advantage of increasing the gain rather than diminishing it. Some operators prefer it to the use of a crystal for obtaining high selectivity by reason of the absence of the "ringing" effect commonly experienced with crystal filters having very narrow band-width.

Mechanical Method.—Another method uses a device for mechanically varying the degree of coupling between the primary and secondary transformer windings. The resonant frequency of the transformer may change as the coupling is varied, and care should be exercised where there are several such transformers, as in a multi-stage I.F. amplifier, to ensure that all of them remain in tune together.

Tertiary Winding.—Another very satisfactory method uses an adjustable

electrical coupling element. This may be either inductive or capacitive. A typical circuit is shown in Fig. 66. A two-stage amplifier provided with three transformers of this type operating at 560 kc/s. can be made to have the characteristics shown in Fig. 67.

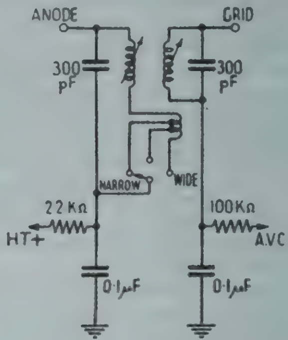


Fig. 66.
A tertiary winding may be used in an I.F. transformer to provide adjustable selectivity.

Crystal Filters

By incorporating a crystal filter in an I.F. amplifier, the selectivity can be very greatly enhanced. Whereas a normal tuned circuit resonating at about 450 kc/s. cannot be made to have a Q-value higher than 200–400, a quartz crystal at the same frequency will have a Q of about 20,000: the crystal, which behaves as the equivalent of a resonant circuit, thus has a correspondingly higher degree of selectivity.

The equivalent circuit of a crystal is shown in Fig. 68. The capacitance C_0 represents the capacitance between the electrodes together with that of the holder and the circuit connections, and L, C and R represent inductance, capacitance and resistance corresponding to the vibrational characteristics of the crystal. A circuit of this from has two resonant frequencies; (*a*) a series-resonance frequency, f_s, determined by L and C, and (*b*) a parallel-resonance frequency, f_p, determined by C_0 together with L and C. The value of f_p can be varied over a small range by adding an adjustable capacitance in parallel with C_0. At the series-resonance frequency, f_s, the crystal has a low impedance, and at the parallel-resonance frequency, f_p, which is only very slightly higher than f_s, the crystal has a high impedance. It is by reason of these properties that crystal filters can be made to have extremely high selectivity and to permit the rejection of an unwanted signal occurring near the wanted frequency. A pair of crystals suitably matched can be made to provide a band-pass filter with a remarkably steep-sided characteristic.

A typical crystal filter circuit incorporated as part of an I.F. amplifier is shown in Fig. 69. The I.F. transformer L_1L_2 has a centre-tapped low-impedance secondary, the output being taken through the crystal

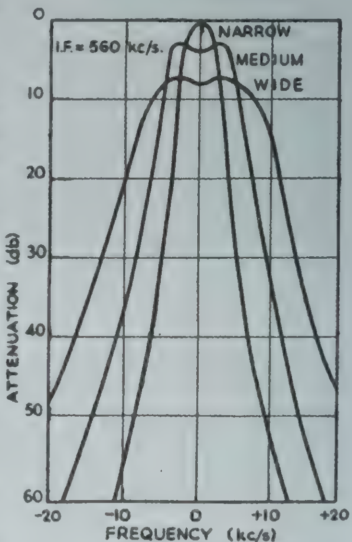


Fig. 67
The selectivity characteristics obtainable with an adjustable tertiary winding used in an I.F. transformer as shown in Fig. 66.

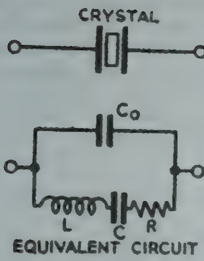


Fig. 68.
The equivalent circuit of a quartz crystal.

X and the balancing condenser C. The inductance L_3 is, in effect, a voltage step-up transformer having a suitable impedance ratio for driving the grid of the following amplifier.

The condenser C is adjusted so that it is approximately equal to the effective shunt capacitance of the crystal holder. This is necessary in order to balance out the R.F. current which passes through the crystal irrespective of its vibrational characteristics, *i.e.* by reason of its resemblance to an ordinary condenser. The circuit thus becomes a simple bridge circuit which is balanced for all frequencies other than the frequency at which the crystal behaves differently from a simple condenser.

If the incoming signal produces an I.F. voltage having exactly the same frequency as the series-resonance frequency of the crystal, the crystal behaves as a very low resistance and the bridge ceases to be balanced: energy is then effectively transferred to the succeeding amplifier.

If the incoming signal produces an I.F. voltage having exactly the same frequency as the series-resonance frequency of the crystal, the crystal behaves as a very low resistance and the bridge ceases to be balanced: energy is then effectively transferred to the succeeding amplifier.

If the frequency differs slightly from the series-resonance frequency, the crystal approximates to a condenser and behaves as a high impedance—much higher than the transformer windings to which it is connected—and therefore very little energy is passed on to the amplifier. This constitutes a high-selectivity frequency-filter, the width of the pass-band depending on the relative impedances of the crystal and the tuned circuits.

By setting the condenser C_2 so that the transformer secondary L_2 is detuned from the crystal frequency, it can be made to have a relatively low impedance: in comparison with this the crystal impedance appears even greater and the selectivity is increased. When L_2 is tuned to resonance, it presents its maximum impedance, in comparison with which the crystal impedance is not so much greater, and the selectivity is then at its minimum. The condenser C_2 can thus be used to control the degree of selectivity. It should be noted that the detuning unavoidably results in a loss of signal strength, but this loss is never serious and there is usually a reserve of gain in the I.F. amplifier.

When the balancing condenser C (sometimes known as the *phasing condenser*) has exactly the same capacitance as that across the crystal (*i.e.* the capacitance of the holder and the associated capacitances), the selectivity

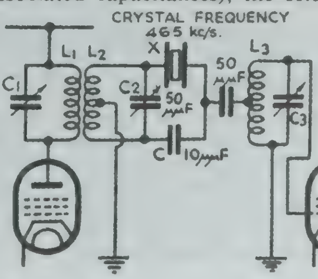


Fig. 69.
A simple crystal filter circuit suitable for I.F. amplifiers operating at frequencies in the region of 465 kc/s. The condenser C is usually made variable so as to provide an adjustable phase control.

curve is symmetrical about its mid-point as shown in Fig. 70A. A slight variation from the balancing setting of C causes the curve to become asymmetrical: see Fig. 70B. The sharp dip in the response which occurs on one side of the peak is due to the parallel-resonance of the crystal: the frequency

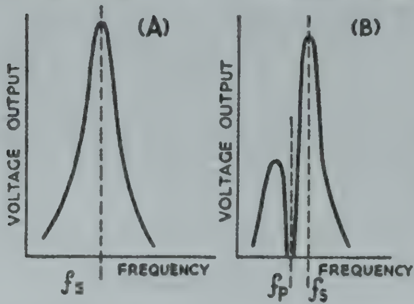


Fig. 70.
When the crystal filter is exactly balanced the selectivity curve is symmetrical (A). When the balancing condenser is varied the curve becomes asymmetrical (B).

at which this parallel-resonance occurs is always very close to the series-resonance frequency, but its actual value depends on the effective parallel capacitance. It is this effective parallel capacitance which is altered by varying the phasing condenser C. The frequency at which the sharp dip occurs can thus be varied by means of the phasing condenser.

This feature is useful in greatly reducing the intensity of an unwanted heterodyne arising from a carrier close to the frequency of the wanted signal. The sharp parallel-resonance dip is sometimes known as the *rejection notch*.

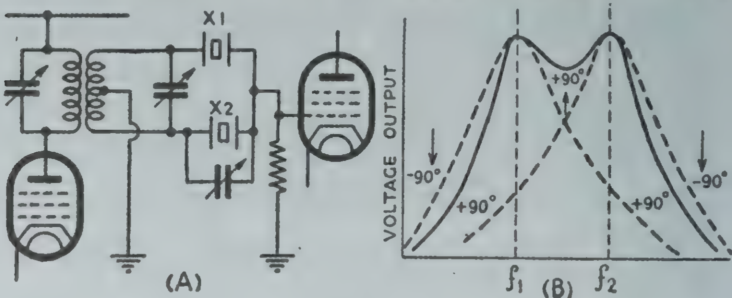


Fig. 71.
A band-pass crystal filter using a pair of crystals. The combined response curve has steeper sides than the curve of either crystal considered separately.

Crystal Band-pass Filters

By using a pair of crystals having resonant frequencies differing by, say, 500 c/s., the filter can be made to have a broad-topped steep-sided characteristic. The second crystal can be added to the circuit shown in Fig. 69 by inserting it as a replacement of the condenser C: see Fig. 71A. In general a small shunting condenser will be necessary to ensure the required balance, and it may have to be connected across the crystal X_1 rather than X_2, or *vice versa*, according to their respective self-capacitances.

The combined resonance curve of the dual crystals is shown in Fig. 71B. Since the voltages across the two crystals are 180° out of phase, the combined response is not a simple addition of the two separate responses, except between the peaks where the voltages are in phase. The result is a close approximation to the ideal filter for receiving telephony, the pass-band being slightly wider than the frequency difference between the crystals and the cut-off at the sides being sharper than that obtained with either of the crystals used singly.

Single-signal Reception

In a C.W. receiver where the selectivity is relatively poor, there is no appreciable difference in signal strength obtained on the two sides of the zero-beat frequency. If, however, the selectivity is increased to a high order

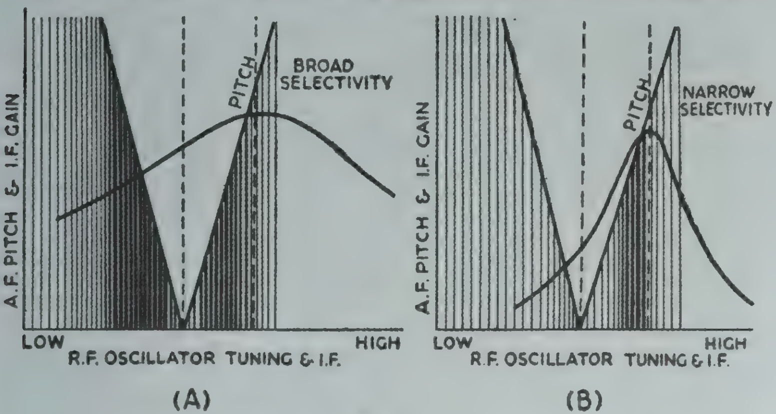


Fig. 72.

In (A) the signal is almost equally strong on both sides of zero beat, whereas in (B) the signal is almost inaudible on the "low" side of zero beat. Where the selectivity is sufficiently sharp to produce a "single signal," as in (B), the following relationships exist.

(1) If L.O. > S.F.—
Increasing L.O. frequency increases I.F. and if B.F.O. < I.F. beat rises.
Increasing L.O. frequency increases I.F. and if B.F.O. > I.F. beat falls.

(2) If L.O. < S.F.—
Increasing L.O. frequency decreases I.F. and if B.F.O. < I.F. beat falls.
Increasing L.O. frequency decreases I.F. and if B.F.O. > I.F. beat rises.

such that a difference in tuning of about 2 kc/s. causes a change in signal voltage of, say, 10 db, it will be found that the signal is much stronger on one side of zero-beat than on the other. If the selectivity is high enough, the signal on the weaker side may be made so weak as to be practically inaudible, while the signal on the stronger side remains strong or is even of greater amplitude. This condition is known as *single-signal reception*.

It is hardly possible to obtain this condition in a straight receiver owing to the inherently low selectivity, but it is readily produced in a superhet because of the high selectivity attainable in the I.F. amplifier. It is more readily obtainable as the frequency chosen for the I.F. is made lower and as the number of I.F. tuned circuits is increased. The use of positive feedback or a

crystal filter will usually raise the selectivity to a peak high enough to produce a single-signal condition.

The strong signal may appear on one side or other of zero-beat according to whether the B.F.O. is tuned on the high or low side of the I.F. resonant frequency and whether the local (R.F.) oscillator is tuned above or below the signal frequency. A study of the single-signal principle shows how it is possible to differentiate between true signals and image signals by observing whether or not the signal appears on the correct side of zero-beat. This can best be explained graphically: see Fig. 72. The illustration shows the simultaneous relationship between the audio pitch of the beat note and its intensity and the dependence of this relationship on the tuning of the local oscillator. In Fig. 72A, the selectivity is of a low order as shown by the broad I.F. resonance curve. The intensity of the beat note varies little with its pitch and is practically the same on both sides of zero-beat. In Fig. 72B, the I.F. selectivity curve is much narrower and if the setting of the tuning control is such that the central zero-beat of the I.F. signal is not coincident with the resonance peak of the I.F. amplifier, the signal will be stronger on one side of the centre zero-beat than on the other.

In most superhets, the R.F. oscillator is higher in frequency than the signal: therefore, if the B.F.O. is set on the *low* frequency side of the I.F. centre frequency, the beat note will be heard much more strongly when the main tuning control is set on the *high* side of the carrier frequency. If the selectivity is sharp enough, the signal on the lower side will be inaudible. Thus, when these conditions exist the heterodyne beat note will *rise* in pitch when the tuning control is moved *higher* in frequency. This will also occur if the R.F. oscillator operates on the *lower* side of the carrier frequency and if the B.F.O. is set on the *high* side of the I.F. resonance frequency. In the other two possible combinations, the pitch of the beat note will *fall* when the tuning control is moved higher in frequency.

As a consequence of this, in the single-signal condition, a signal which is not a true signal but actually an image will appear on the " wrong " side of zero-beat; *i.e.* if the adjustments are such that all true signals *rise* in pitch as the tuning control is moved in one direction, an image will *fall* in pitch. This is illustrated in Fig. 73 which shows a typical group of C.W. carriers at

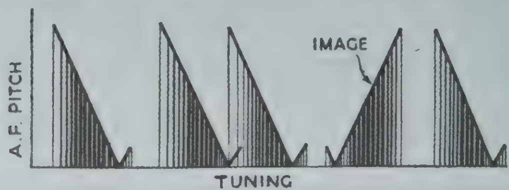


Fig. 73.
In a single-signal receiver an image is easily distinguished from true signals by the reversed variation of pitch.

various frequencies and an image that occurs amongst them. As the tuning control is moved steadily in one direction, the ear will instantly detect the image by the reversed variation in pitch. If the selectivity is not high enough to make the signal appreciably stronger on one side of zero-beat than on the other, it will obviously not be possible to detect an image in this way.

The degree of selectivity required for single-signal reception is too high to permit good speech quality (by amplitude modulation). The frequency

pass-band may be of the order of 300 c/s., whereas a range of audio frequencies from 300 to 2,000 c/s. is considered necessary for "commercial" quality.

I.F. Amplifier Gain

Most of the gain in a superhet takes place in the I.F. amplifier. A typical pentode stage, as shown in Fig. 74, may produce a voltage gain of 200.

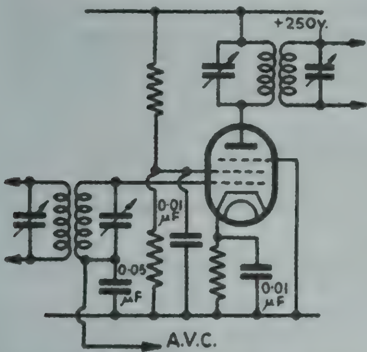


Fig. 74.

A typical single-stage I.F. amplifier. The potential divider which supplies the screen potential and the cathode bias resistance must be selected to suit the type of valve used.

The figure varies considerably according to the type of valve, the voltages at which it is operated and the type of inter-stage coupling. Almost any type of R.F. pentode may be used in an I.F. amplifier, but it should be remembered (i) that the frequency of operation is relatively low and therefore there is no point in using a special V.H.F. type, and (ii) that the last stage of an I.F. amplifier may have to handle a rather large grid-voltage swing and the type of valve chosen should be capable of handling this swing without over-driving and grid. Some of the favourite valves for I.F. amplifiers are—

6D6 6K7 6S7 6SS7
12SK7 EF39 KTW61

In low-priced receivers, there is usually no separate external control of the gain in the I.F. amplifier, but this is a feature which could well be included when possible. The most convenient form of gain is a variable cathode bias resistance (such as a conventional 3-watt potentiometer). Another form is a potentiometer control of the screen potential: see Figs. 50 and 53.

A.V.C.

Automatic volume control, so-called because it automatically adjusts the "volume" of sound from the headphones or loudspeaker to a predetermined level, is in effect the same as A.G.C. (automatic gain control). As applied to ordinary communications receivers, the two terms are synonymous, but in television receivers the term A.G.C. is obviously preferable.

The principle used in a simple A.V.C. system is as follows: the mean carrier signal level, averaged over a suitable period of time (usually of the order of one second), is used to control the gain of the amplifier in such a way that the gain is reduced as the signal level rises, and *vice versa*.

One essential requirement of all A.V.C. systems is an ample reserve of gain. If the receiver is operating near to its maximum gain when the signal level is at its mean value or somewhat above it, the gain cannot be increased

sufficiently to produce constant output when the signal level falls. For this reason, A.V.C. is not usually found in straight receivers.

The most convenient method of gain control is the regulation of the grid bias, the bias voltage being derived from a diode rectifier fed from the I.F. amplifier. Fig. 75 is a block diagram showing how this can be incorporated into a conventional type of superhet. The A.V.C. voltage is shown applied

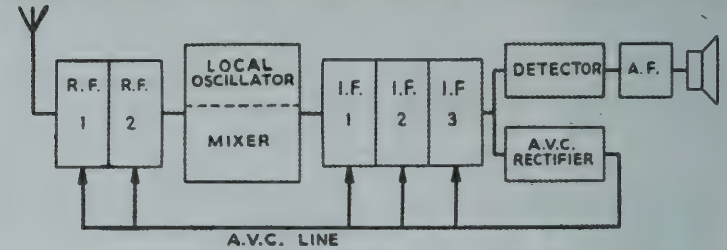


Fig. 75.
A.V.C. may be applied to all R.F. and I.F. amplifier stages. It is preferable not to apply it to the mixer-oscillator in order to avoid the risk of frequency-shift.

to all stages where R.F. or I.F. signals are amplified. Thus the gain of every amplifier stage (prior to the detector) is controlled by the signal, and the overall effect is to give a sensitive control of amplitude, *i.e.* a high degree of amplitude stability. It is preferable not to apply the A.V.C. bias voltage to the mixer because there is a risk that the varying anode current in that valve may affect the oscillator frequency, and the tuning adjustment would then be found to vary slightly with the signal intensity.

The R.F. and I.F. amplifier valves should be of the variable-mu type to ensure good A.V.C. characteristics.

The diode used for supplying the A.V.C. voltage can be of the same kind as that used for demodulation (*i.e.* the detector). The output from the A.V.C. diode is smoothed in a resistance-capacitance filter having a suitable time-constant: for medium-wave broadcast signals the time may be about 0·2–0·5 second, and for amateur-bands receivers it may be about 0·1 second.

A typical circuit arrangement is shown in Fig. 76. Here the A.V.C. diode and the signal diode are combined in one valve together with the triode

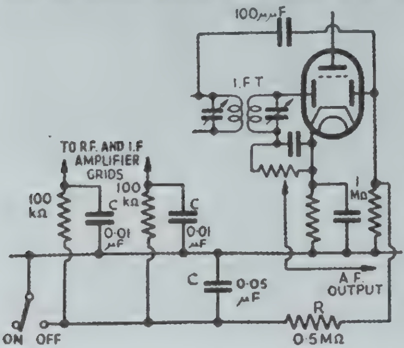


Fig. 76.
An A.V.C. circuit suitable for a typical superhet. To lengthen the time constant, the values of the condensers C and the resistances R should be increased.

A.F. amplifier (the well-known double-diode-triode). This is merely for convenience and economy in construction. Technically, the three functions are quite separate and distinct.

Any R.F. current which reaches the A.V.C. diode will be rectified and converted into A.V.C. bias. Thus, noise and interference will reduce the gain in proportion to their amplitude. Since there is always an appreciable background of noise in a high-gain amplifier, the total gain will be prevented from reaching its maximum. It is therefore desirable to render the A.V.C. inoperative until the incoming signal is fairly strong or at least above a certain minimum level: in other words, it is desirable to delay the effect of A.V.C. (in *level*—not in *time*, of course). This is achieved by introducing a small positive bias to oppose the negative bias produced in the A.V.C. rectifier circuit. A suitable delayed A.V.C. circuit arrangement is shown in Fig. 77.

In most C.W. superhets, the B.F.O. injects a voltage into the I.F. and detector circuits which is usually sufficient to produce a large A.V.C. voltage. This means that it is not practicable to use A.V.C. when the B.F.O. is switched on, as the sensitivity of the receiver is very greatly reduced. By weakening

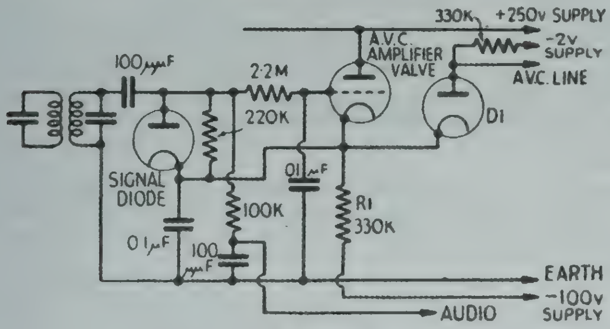


Fig. 77.
A circuit for providing amplified A.V.C.

the coupling of the B.F.O. to the detector, or by using adequate screening to ensure that the B.F.O. voltage reaches nothing but the signal detector, it is possible to reduce the A.V.C. voltage arising from it so far as to cause only a trifling decrease in sensitivity while still providing a sufficiently strong beat note with incoming signals. It is also necessary to lengthen the time-constant of the A.V.C. filter so that the A.V.C. bias does not fall appreciably in the intervals between the Morse characters.

Generally speaking, the use of A.V.C. for C.W. reception is not much required, for the aural effect of variations in strength of C.W. signals is less objectionable than variations in loudness of telephony signals.

The Tuning Eye

The miniature cathode-ray tube known as the *tuning eye* (or *magic eye*) is a very useful and convenient indicator for showing the amplitude of the R.F. signals fed into the detector. This is its most popular application, but it is also used as an indicator in several forms of auxiliary equipment, such as frequency meters.

The luminous pattern which appears on the circular screen at the top of the tube is dependent on the grid potential of the triode incorporated with the cathode-ray projection electrodes. A common form of pattern is a V-shadow, illustrated in Fig. 78, which also shows a typical circuit arrangement for a tuning eye in a superhet.

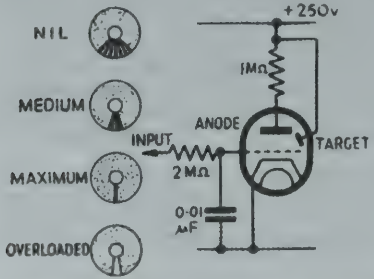


Fig. 78.
A typical tuning-eye circuit. The input may be taken from the A.V.C. line.

The tuning eye is particularly helpful when the circuits of a superhet are being aligned since it gives a direct indication of the signal intensity at the output end of the I.F. amplifier. It will usually be necessary to switch-off the B.F.O., for the B.F.O. injection voltage may be sufficient to close the eye and thus mask the effect of the tuning adjustments.

S-Meters

Although the so-called S-meter has achieved rather widespread popularity as a means of measuring signal strength, it is a moot point whether it has any *real* value. The circuits associated with it may easily be misadjusted or the indication which it gives may be misinterpreted.

The usual arrangement does no more than indicate the relative strength of the incoming carrier and takes no account of the amplitude of modulation. This means that there is no direct relation between the sound intensity of a telephony transmission and the reading given by an S-meter. The adjustment of the aerial coupling and the R.F. amplifier will also affect the behaviour of the meter.

If these limitations are borne in mind, the inclusion of an S-meter may be justified. A typical circuit is given in Fig. 79.

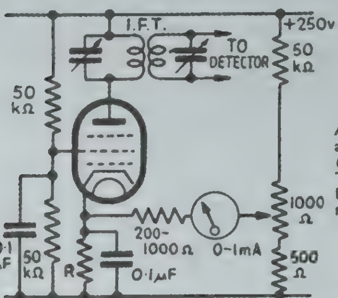


Fig. 79.
A popular form of S-meter circuit incorporated in a typical I.F. amplifier. Any variable-mu valve (such as a 6K7, EF39 or KTW61) may be used. The values of the resistances shown may have to be altered to obtain the best performance. The normal cathode bias resistance should be used for R.

The B.F.O.

The frequency stability of the B.F.O. is more important than is sometimes appreciated. Any variation will show up immediately in a corresponding variation in the pitch of the beat note. It is therefore desirable to feed the B.F.O. from a voltage-stabilised supply (the same one as is used for the R.F. local oscillator can be employed for this purpose) and to choose a circuit which

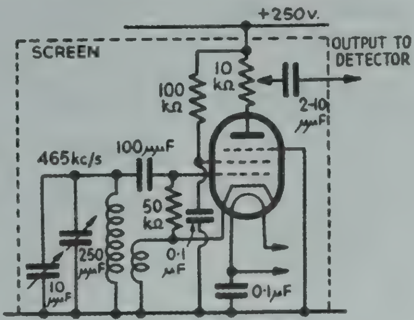


Fig. 80.

A typical B.F.O. circuit for medium frequencies (e.g. 465 kc/s.). A triode may be used instead of the pentode shown here if the high stability of the E.C.O. type of circuit is not considered necessary.

is inherently stable. A conventional E.C.O. type of circuit is usually found satisfactory. The valve should be operated under conditions of light load in order to minimise the effect of thermal drift. A suitable circuit is shown in Fig. 80.

The injection of the B.F.O. voltage into the detector circuit is best made to the detector through a small condenser of 2–10 μμF.

Noise Limiters

A distinction must be drawn between two main classes of noise: (i) the continuous "rushing" sound which is present as a background in all high-gain receivers, and (ii) the sharply-peaked intermittent crackle associated with such sources as car ignition systems. It is only the latter type of noise that can be suppressed by the present-day noise limiter circuits. The reason for this will be apparent from the following explanation of the basic method used in these circuits.

The ignition type of noise is characterised by an exceptionally high peak amplitude and a very short period of duration. The modulation frequencies which constitute this noise are, therefore, mostly rather high in the A.F. spectrum and the corresponding R.F. band-width is rather large. In simple terms, the R.F. noise pulses are allowed to be amplified together with the signal and then rectified so as to provide a biasing pulse which can be fed back to the amplifier and thus momentarily reduce the amplifier gain. During this pulse period, the incoming signal will, of course, be reduced in amplitude, but roughly speaking, the effect is more of a levelling-out of the general sound output than a severe "silencing" of the signal by the A.V.C. action of the noise pulses; moreover, the ear is disturbed less by "holes of silence" in the signal than by superimposed peaks of noise energy. The result is a great improvement in the intelligibility of the signal.

For the noise-suppressing action to be effective, it is important to include all the high-frequency components in the noise pulse. If the I.F. amplifier is adjusted to give narrow-band selectivity, the sharpness and the peak

intensity of the noise pulses will be diminished and the A.V.C action correspondingly reduced. A troublesome amount of noise energy will still be present in the sound output of the receiver. Therefore, the selectivity should be reasonably broad if effective noise limiting is required. In the more elaborate systems which have been devised, it is possible to operate the receiver at high selectivity and still have an effective noise-limiting action.

Sometimes, when a noise limiter has been added to an existing receiver, the results have fallen short of expectations. Possibly much of the disappointment can be attributed to a failure to appreciate the nature of the problem. Provided that careful attention is paid to the band-width requirements and to the time-constants in the noise rectifier, a limiter can be added to any orthodox superhet with the assurance that it will justify itself.

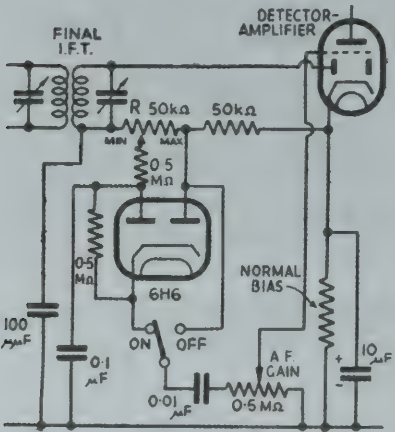


Fig. 81.

An effective noise limiter connected in the detector circuit of a typical superhet. The components should be arranged so as to avoid unwanted capacitances as far as possible, since the operation of the circuit depends on the preservation of the sharply-peaked waveform of the noise pulses. It may also be worth while to screen the components and the connecting leads in order to prevent hum pick-up.

Several different circuit arrangements have been proposed, and various operators have their own favourites. There is space here only for one typical circuit: see Fig. 81. The potentiometer *R* provides an adjustment for determining the modulation percentage threshold at which the limiting actions begins. When the potentiometer *R* is set to Min. the limiting action begins at 100 per cent. modulation, and as the setting is advanced towards Max. the threshold level is reduced. The limiting action is automatically adjusted to the instantaneous carrier level.

The circuit is rather susceptible to stray fields, and it is preferable to fit a screen round the components in order to minimise A.C. hum pick-up. Care should be taken to avoid introducing unwanted by-pass capacitances.

R.F. Amplification in the Superhet

In many of the cheaper and simpler forms of superhet the signal is fed directly from the aerial into the mixer. The disadvantages of this arrangement are chiefly (i) poor discrimination against image signals, and (ii) poor signal/noise ratio. Both of these disadvantages can be largely overcome by the use of a stage of R.F. amplification in front of the mixer valve. An even further improvement can be obtained by using two R.F. stages, and this is

often considered worth while. Three stages have been used in some receivers but the extra trouble in design and adjustment weigh rather heavily against the possibility of improved performance.

Any normal R.F. amplifier (see Chapter 4) can be used in front of the mixer stage, and if A.V.C. is required a variable-mu type of valve will naturally be preferable.

The selectivity of the mixer grid circuit will be much higher when it is preceded by an R.F. stage, and consequently it will be necessary to use more care in the adjustment of the frequency-tracking components.

A.F. Amplification in the Superhet

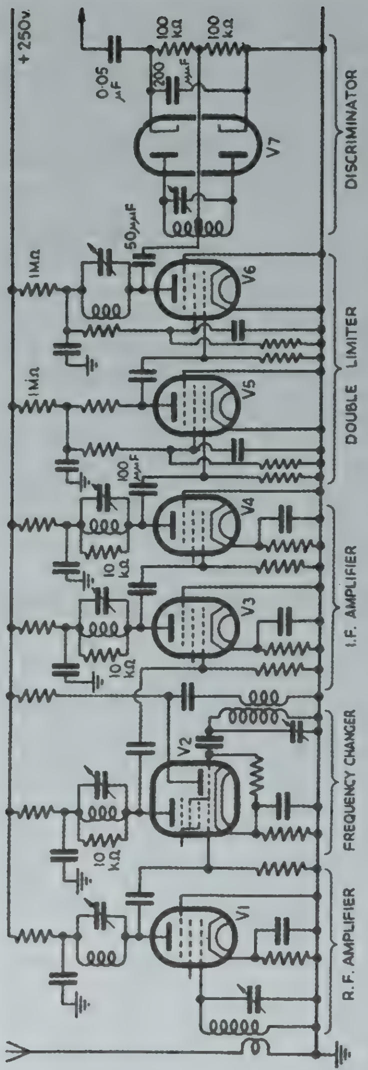
If the superhet is designed primarily for C.W. reception and if the selectivity of the I.F. amplifier limits the pass-band to, say, $\pm$ 3 kc/s. (which is quite adequate for medium-quality speech), it will obviously be unnecessary and wasteful to provide a high-quality A.F. amplifier. Moreover, for C.W. reception, the amount of A.F. power required from the output stage need not exceed about 1 watt. This means that the A.F. amplifier in a C.W. superhet can well consist of a small triode or tetrode having a stage-gain of perhaps 10–30 and an anode dissipation of less than 4 watts.

A high-quality A.F. amplifier may be used with a superhet if the I.F. amplifier has a suitably broad pass-band of, say, $\pm$ 8 kc/s. or more. The loudspeaker should not be mounted on the tuning chassis or be placed close to the tuning condensers owing to the risk of vibration of the condenser vanes and the consequent frequency modulation of the associated circuits: this acoustic feedback is apparent as a howl of constant pitch and occurs only when the receiver is tuned to a carrier.

CHAPTER 6 FREQUENCY-MODULATION RECEIVERS

ALTHOUGH the advantages of frequency-modulation, as a system of transmission and reception, are generally considered to be most attractive when the carrier frequency is high, *i.e.* where wide-band transmission can be tolerated without appreciable risk of interfering with other transmissions, there are nevertheless certain attractions in the use of F.M. on the lower frequency bands, *e.g.* 14 Mc/s. Since the majority of receivers in use at the present time are designed for *amplitude* modulation, it follows that any telephony transmission may be capable of causing interference (*e.g.* by " break-through " into the I.F. amplifier, by " shock-excitation " of the input circuit, or as an image signal in the case of a superhet receiver) provided that the transmitter is using amplitude modulation. If the transmitter is using *frequency* modulation, there should be no interference whatever since the ordinary receiver is not designed to be sensitive to this form of modulation. Strictly speaking, there remains a slight risk where the receiver contains circuits of very high selectivity, for in effect such a receiver is an F.M. receiver besides being an A.M. receiver.

In all F.M. receivers there must be a high degree of gain in order to produce a strong enough signal under all possible conditions to saturate the limiter. It is hardly practicable to achieve sufficient gain in a straight receiver on frequencies for which F.M. is considered worth while, and a superhet is therefore invariably used. The width of the pass-band in the I.F. amplifier must be at least as great as the full frequency deviation used by the transmitter; otherwise the amplitude peaks will be restricted and speech



modulation will be distorted. As long as there are transmitters in operation using various degrees of deviation, it is desirable to design the receiver so as to have the maximum band-width likely to be required. In amateur practice this figure may be fixed at $\pm$ 15 kc/s.

The R.F. and mixer stages in the usual kind of superhet will almost certainly have sufficiently wide-band characteristics and no modification need be considered. The I.F. amplifier should provide substantial amplification but the tuned circuits should have low *Q*-values. Instead of the conventional type of I.F. transformer having two windings loosely coupled, the I.F. amplifier couplings in an F.M. receiver may be single-tuned circuits, possibly shunted with fixed resistances to broaden the resonance. Because of the larger band-width, the stage gain is considerably lower than is usually present in highly-selective amplifiers, and therefore in F.M. receivers there should be at least two I.F. stages, each being operated under conditions giving the maximum gain consistent with the necessary band-width. Any of the popular types of R.F. tetrode or pentode, or any television-type of amplifier valve will be suitable, since a high input impedance is not an important requirement.

After the amplifier proper, one more broadly tuned stage is required, not for amplifying but to limit all incoming signals to a predetermined amount in order to remove all forms of amplitude modulation. This limiter consists of a tetrode or pentode operated with low anode and screen voltages so that it is easily saturated by even the weaker signals. A cascade limiter, using two such stages, gives improved limiting action over a wide range of input signal intensity.

Fig. 82.
A suggested circuit diagram of an F.M. receiver. The output may be fed into any normal A.F. amplifier. To avoid congestion in the diagram, the screen circuits of the first four valves have been omitted: they should be designed in accordance with the usual recommendations for the types of valve selected.

The values of the grid condensers and resistances in the limiter are chosen so that while the first half of the limiter operates best on rapid pulses the second half operates best on slower variations of amplitude.

A double diode with separate cathodes is used in the discriminator (see Fig. 22). This gives an A.F. output the amplitude of which is proportional to the amount of frequency deviation.

The complete circuit arrangement from the aerial to the discriminator output is shown in Fig. 82. Any conventional form of A.F. amplifier may be used to follow the discriminator: as a rough guide to the amount of A.F. gain required, it may be assumed that the output from the discriminator is of the same order as that ordinarily obtained from the diode detector in an A.M.-type of superhet.

CHAPTER 7 CONSTRUCTION

PERHAPS it is true that only the minority of amateurs actually build their own receivers nowadays. This is not greatly to be deplored, for the construction of a modern communications receiver can absorb a prohibitive amount of time. On the other hand, much useful knowledge can be gained in the process of designing and constructing a receiver.

Few amateurs are content for long to leave things as they are. Even the best factory-made equipment becomes liable to some modification or other in the hands of the average amateur—such as the changing of the tuning arrangements, or the fitting of an S-meter or a noise limiter. Ex-Service receivers usually require constructional modifications if they are to be entirely suitable for amateur purposes. A knowledge of the important features of construction and of tne pitfalls to be avoided will often preclude disappointment and prevent waste of effort.

From time to time, a factory-made receiver may require attention to cure faults which have developed in its various components or readjustments may be necessary to restore its normal performance. An experienced constructor finds such a task much easier than the amateur who has hitherto ignored the constructional side of receiver design.

A first-class factory-made receiver is unavoidably expensive. This does not mean that the buyer gets poor value for his money. Often the value is remarkably good, but far too much work must be put into it for it to be marketed at a really low price. The home-construction of a receiver therefore continues as a strong attraction to the amateur who must limit his expenditure. He can be assured that, with care, he should be able to produce a receiver at least as good as the receivers offered by the manufacturers in the top-price class.

It frequently happens that an amateur's requirements cannot be met exactly by any receiver available on the market and this often constitutes a sufficiently strong incentive for home-construction. The design may be simple or elaborate, and the cost may be trivial or extravagant: the wholesome satisfaction to be derived from such specialised effort is the complete answer to the question—Is it worth while?

The Tuning Mechanism

The heart of a communications receiver, in the constructional sense, is the tuning mechanism. This includes the tuning scale and the tuning knob.

The dimensions and the physical characteristics of these two items are overwhelmingly important to the success of the receiver. Here are the chief requirements:

(1) The knob should be about 2 in. in diameter and the spindle should be horizontal at a height of about 4 in. above the desk. It should not have sharp edges or corners, nor should it be very deeply fluted.
(2) The friction on the spindle should be sufficient to prevent unintentional slipping but not so great as to prevent easy adjustments of a fraction of a degree by light finger pressure. There should be no suggestion of springiness or " rubbery " feeling and no backlash.
(3) The scale should be as large as possible and preferably illuminated from the rear. A uniformly-lit, well-figured dial is a great help towards good operating.
(4) There should be no parallax: this implies that the pointer or index mark must be very close to the plane of the dial or scale.
(5) It is preferable to have a fixed index mark and to move the dial or scale so that the figures are carried past the index mark rather than a rotating pointer moving over a fixed semi-circular scale, but a good alternative is a fixed horizontal linear scale with a sliding index mark moving horizontally. The plane in which this scale lies should be tilted so that it can be viewed squarely from the normal operating position. In most receivers the dials can be viewed squarely only when the operator's chin is resting on the desk. If the constructional difficulties are too great, the whole receiver may be tilted at the necessary angle.

Certain other controls should be located near the tuning knob. These are—
Gain controls: R.F., A.F. (and I.F. if provided).
C W./Phone switch (*i.e.* B.F.O. on/off switch).
Selectivity control.

The less-frequently used controls, such as the A.V.C. on/off switch and the tone control can be located farther away from the tuning knob without serious loss of operating convenience.

In mains-energised receivers, it is often better to build the power supply as a separate unit. This reduces the risk of frequency-drift of a superhet oscillator due to the temperature rise, and it also reduces the risk of mains hum. The weight of the mains transformer and the smoothing chokes does not have to be carried on the receiver chassis, and if necessary the power unit can easily be disconnected from the receiver and used for supplying other equipment.

Sheet-metal Work

The conventional rectangular-dish type of chassis is reasonably well suited to amateur receiver construction, although there is some overcrowding of the components below the deck if the chassis is too small. Excessive depth in the chassis tends to make the assembly rather difficult.

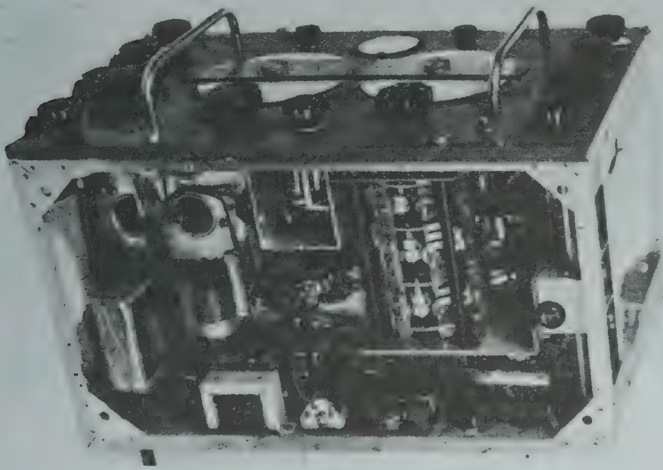
Steel is favoured by the manufacturers for chassis construction, but for amateur purposes aluminium is much to be preferred as it is a good deal easier to work. The lesser rigidity of an aluminium chassis can easily be overcome by fitting strengthening brackets or partitions (which may also help in the internal screening).

The cutting of large holes, such as are required for valve sockets and drop-in power transformers, often presents a difficult problem to the amateur constructor who has only a few simple tools. However, sheet aluminium

can be cut quite quickly with an ordinary wood-cutting fretsaw, using a fine-toothed blade: the cutting is made even easier if the work is lubricated with turpentine. This is a practicable method for thicknesses as great as $\frac{1}{8}$–$\frac{3}{16}$ in.

Aluminium-faced plywood is sometimes more suitable than a dished form of metal chassis, *e.g.* when the sides of the box-form would be unnecessary or would make the assembly and the wiring-up too difficult.

Perhaps the most formidable part of sheet-metal work for home constructors is the bending of long straight edges as in the making of a chassis. If a large vice (about 4 in.) is available, it is possible to make a fairly sharp bend by clamping part of the metal sheet between two pieces of angle iron of suitable length.



(Courtesy *Denco Ltd.*)

An underside view of a factory-built superhet. Mechanical bandspread is used and the frequency coverage is 175 kc/s.–36 Mc/s. The turret switching assembly can be seen on the right.

Layout of Components

In arranging the layout of a receiver it is wise to keep the grid circuits of amplifier valves well away from any part connected to the A.C. supply in order to avoid mains hum. The grid circuits of the detector and the first A.F. stage demand special care. Electrostatic screening may be necessary.

An A.F. transformer is liable to pick up hum from a mains transformer, and since it is difficult to provide effective magnetic screening, the simplest solution to the problem is usually the adequate separation of the transformers or possibly changing the orientation of the cores so that the coupling between the magnetic fields is reduced to a minimum.

Wiring-up

In wiring-up, it is very important to keep all R.F. leads as short as possible.

Many troubles, *e.g.* unwanted oscillation, erratic variation of frequency in oscillator tuning, poor performance and distortion, are often due to excessively long R.F. leads.

Such troubles arise from the unintentional inductance and capacitance introduced into the circuit. The dangers are reduced if the R.F. leads are make of thick wire (*e.g.* 16 S.W.G.) and if they are well separated from the other wiring and the components and the chassis.

Especially important are the earth-return leads: all such leads belonging to a particular amplifier stage should be connected to the same point on the chassis. Even in A.F. amplifiers it is desirable to keep all leads short, for otherwise there is a risk of picking up R.F. energy from the transmitter.

However convenient it may appear to be, an earth-return should never be made to the panel itself. To do this would cause R.F. currents to circulate in the panel, and this would give rise to irritating hand-capacitance effects.

Every soldered joint must be absolutely reliable. Faulty performance can easily be caused by a " dry " joint, but such joints are sometimes difficult to find. Clean surfaces are vital to the making of satisfactory joints and sufficient time must be allowed for all the parts of the joint to become thoroughly hot. In conflict with this, unfortunately, is the advice to use as little heat as possible where the insulation is of low melting-point material: extra care is then obviously required.

It is often convenient to connect small resistances and condensers in the circuit by suspending them in the wiring (*i.e.* by using their own connecting leads to support them). This is a good practice in that it permits the leads to be kept very short, but it is not to be recommended where there is an excessive concentration of components or where there is a risk of unintentional contacts due to possible displacement of the components. Tag boards or strips provide a neat method of mounting, but where they are used care must be taken to avoid long leads, especially in R.F. circuits.

Manufacturers frequently adopt the practice of tying together groups of wires so as to make tight bundles. Undoubtedly this has great advantages

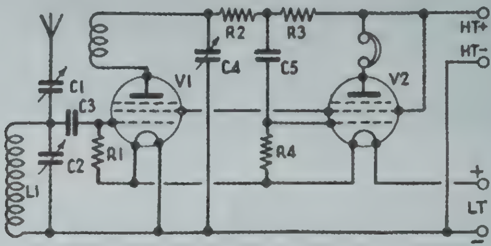


Fig. 83.

A practical circuit diagram of a simple battery-operated receiver. Coil sizes are given in Table II

C1	10 $\mu\mu$F. variable.	R1	2 megohms 2-watt.
C2	100 $\mu\mu$F. variable.*	R2	5,000 ohms ½-watt.
C3	100 $\mu\mu$F. mica.	R3	20,000 ohms ½-watt.
C4	300 $\mu\mu$F. variable.	R4	0·5 megohm ½-watt.
C5	0·01 μF. paper.		

V1 and V2 may be 1C5GT, 1C5G, N14 or other similar pentodes.

* For band-spreading, C2 may be shunted by an additional variable condenser of about 15 $\mu\mu$F.

in large-scale production, but there is no advantage to the individual constructor, and it is a positive nuisance when any work has to be done on the wiring.

In most variable condensers the frame is in contact with the moving vanes and the fixed vanes are carried on insulators. It is therefore preferable to connect the spindle and the frame to the "earthy" side of the tuning circuit. Where the grid bias or the A.V.C. voltage is applied *through* the tuning coil it may be necessary to insulate the frame from the chassis, effectively earthing it through a large by-pass capacitance.

Soldering tags used for making connections to a chassis must be clamped tight. It is unwise to mount them on the screws that are used for fixing such components as paxolin valve holders, for there is a risk of gradual shrinkage of the insulating material with age and the consequent uncertainty of contact.

A Two-valve Receiver

Many of the early records in long-distance communication were made with a simple two-valve receiver comprising a regenerative detector and one A.F. amplifier stage. Sensitivity and signal/noise ratio are both high, and apart from its inferior selectivity, this type of receiver can still hold its own amongst modern superhets.

Band	L_1	Reaction
1·8 Mc/s.	55 turns close wound	7 turns
3·5 Mc/s.	28 turns close wound	4 turns
7 Mc/s.	15 turns spaced to 1¾ in.	3 turns
14 Mc/s.	7 turns spaced to 1¼ in.	2 turns
28 Mc/s.	4 turns spaced to 1¼ in.	1½ turns

TABLE II.

Coil sizes for the 2-valve receiver shown in Fig. 83. All coils are wound with 24 S.W.G. d.c.c. on 1½ in. diameter formers.

A practical circuit diagram of a battery-operated receiver is shown in Fig. 83. The two low-consumption pentodes are capable of being run from a medium-sized 3-volt dry battery for L.T. and an H.T. battery of 45–60 volts. The filaments are connected in series and the anode current consumption need not exceed 5–10 mA.

The coil sizes specified in Table II are for amateur-bands coverage, but with the tuning condenser of 100 μμF., as shown, much wider frequency ranges will be obtainable, and there should be no difficulty in tuning to the various short-wave broadcast bands. The reaction coil must be wound in the same direction as the grid coil, the grid end and the anode end being outermost, and the gap between the two windings being ¼ in. Standard 4-pin plug-in coil formers are well-suited to this receiver.

The old-style breadboard mounting may be used but a metal chassis is recommended. All the components in the left-hand half of the circuit diagram must be kept close together, with the shortest possible leads. The remaining components may be arranged in any convenient position.

In operation, the aerial coupling condenser should be adjusted to the minimum capacitance which gives a satisfactorily strong signal. Excessive coupling lowers the selectivity and increases the risk of interference. The performance of a receiver of this type is greatly marred if "overlap" is present in the regeneration control; some advice on the elimination of this fault is

given in Chapter 4. When the regeneration is adjusted correctly, this receiver will be found surprisingly sensitive, although it should be added that, as with most regenerative detectors, some difficulty may be experienced in achieving high frequency-stability at frequencies above about 25 Mc/s.

A Three-valve Superhet

Even a very simple form of superhet offers much greater selectivity than an elaborate straight receiver. The circuit shown in Fig. 84 represents an economically constructed superhet capable of receiving weak signals through comparatively heavy interference.

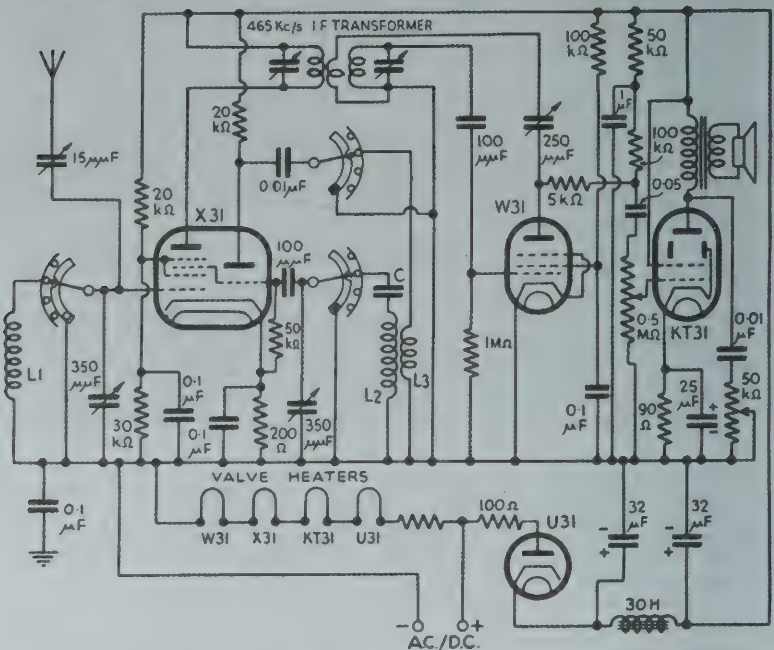


Fig. 84.

A simplified form of superhet using only three valves. It is designed for A.C./D.C. operation and the usual rectifier is included for this purpose. Suggested coil sizes are given in Table III.

For simplicity no R.F. amplifier is used, and to avoid overloading of the mixer grid by strong signals an adjustable aerial coupling condenser is provided; this should be kept at the lowest practicable setting. A combined mixer-oscillator is fully justified in this receiver, and any alternative types (*e.g.* heptodes) could be used in place of the triode-hexode shown.

Only one I.F. transformer is required. This couples the anode circuit of the mixer to the grid of the second valve, which acts as a regenerative leaky-grid detector. Regeneration is effected by the third winding coupled to the secondary of the I.F. transformer. A coil of 5–10 turns should be adequate

for producing self-oscillation. The variable condenser in series with it controls the amount of feedback and will be found extremely useful in sharpening the selectivity when receiving telephony. For C.W. reception, the detector should, of course, be allowed to oscillate. Since this method of producing a heterodyne beat note necessitates some detuning as between the I.F. signal frequency and the detector oscillation frequency, there will be a slight loss of signal strength, and this loss will be greater as the pass-band of the I.F. transformer is made narrower. This loss, however, is not likely to be serious.

Resistance-capacitance coupling is used for driving the output pentode. Moderate loudspeaker reception may be expected under normal conditions but the quality of speech reproduction is limited by the characteristics of the leaky-grid detector.

TABLE III.

Tuning circuit dimensions for Simple Superhet shown in Fig. 84. All coils are close wound on ⅞ in. diameter formers. Spacing between coupled coils is ¼ in.

Coil or Condenser	0·55–1·5 Mc/s.	1·5–4·0 Mc/s.	4–10 Mc/s.	10–25 Mc/s.
L_1	150 turns 30 S.W.G.	36 turns 30 S.W.G.	10 turns 30 S.W.G.	4½ turns 20 S.W.G.
L_2	95 turns 30 S.W.G.	31 turns 30 S.W.G.	9½ turns 30 S.W.G.	4 turns 20 S.W.G.
L_3	15 turns 30 S.W.G.	5 turns 30 S.W.G.	3 turns 30 S.W.G.	2 turns 20 S.W.G.
C	400 μμF.	1,070 μμF.	2,900 μμF.	7,300 μμF.

An I.F. of about 465 kc/s. should prove generally satisfactory, although there will undoubtedly be some image interference on the higher frequency tuning ranges. This will be appreciably lessened if a higher I.F., say 1·6 Mc/s., is used. Since the overall gain from the aerial to the detector circuit is relatively small, no provision is made for A.V.C.

A chassis measuring 10 in. × 8 in. × 2½ in. should provide sufficient mounting space for all the components shown in the diagram. Suggested coil sizes are given in Table III.

The receiver is of the A.C./D.C. type and can be connected to either A.C. or D.C. mains without alteration to the circuit, except for the adjustment of the series resistance to suit the voltage of the supply: even this adjustment should not be necessary if a barretter of appropriate rating is used. With slight modifications, the circuit could be arranged for A.C. only or for batteries.

A High-gain Converter

A converter serves to change the frequency of any incoming signal to some other frequency which can then be fed into any existing receiver (either superhet or straight) tuned to that selected frequency. This principle is

described in Chapter 5 and illustrated in Fig. 63. The simplest converter consists of a combined mixer-oscillator, and an improvement on this is effected by separating the oscillator from the mixer. A very much greater improvement is obtained by using an R.F. amplifier stage, thus making a total of three valves.

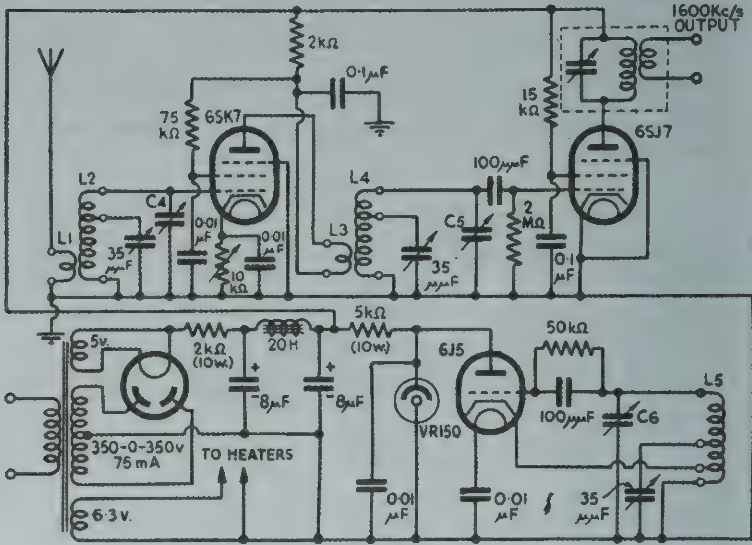


Fig. 85.
A high-gain converter suitable for feeding signals into an existing receiver (superhet or straight type). An I.F. of 1,600 kc/s. is used. Coil sizes are given in Table IV.

The circuit diagram of a converter of this kind is shown in Fig. 85. When coupled to an ordinary superhet or to a good straight receiver with at least one R.F. stage, a highly satisfactory performance may be expected. The degree of image rejection will depend largely on the conversion frequency (see Chapter 5), and for operation on the higher frequency bands an I.F. of 1,600 kc/s. may be recommended.

In the circuit shown here electrical band-spread is provided, the main tuning condensers being connected across only a portion of each coil. Separate trimmers are attached to each coil and these should be individually adjusted to give the best performance. Details of the coils and the trimmers are given in Table IV.

The output circuit comprises the primary winding of a 1,600-kc/s. transformer and a small winding of 10–20 turns very closely coupled to it. The normal secondary winding should be entirely removed. A screened low-impedance lead may be used to connect the output from the converter to the aerial input terminals of the receiver, which are presumed to be arranged for a low-impedance input.

The built-in power supply may be omitted only if the main receiver can be relied upon safely to provide the extra feed current (25–30 mA. at H.T. and approximately 1 A. at 6·3 volts).

TABLE IV.

Tuning circuit dimensions for the High-gain Converter shown in Fig. 85. All coils are wound with 22 S.W.G. enamelled wire on 1½ in. diameter formers. Spacing between coupled coils in ¼ in.

Coil or Condenser	3·5 Mc/s.	7 Mc/s.	14 Mc/s.	28 Mc/s.
L_1	10 turns close wound	8 turns close wound	5 turns close wound	3 turns close wound
L_2	40 turns close wound tapped 26 t.	23 turns spaced to 1½ in. tapped 8 t.	12 turns spaced to 1½ in. tapped 3 t.	6 turns spaced to 1½ in. tapped 2 t.
L_3	14 turns close wound	12 turns close wound	6 turns close wound	3 turns close wound
L_4	40 turns close wound tapped 26 t.	23 turns spaced to 1½ in. tapped 8 t.	12 turns spaced to 1½ in. tapped 3 t.	6 turns spaced to 1½ in. tapped 2 t.
L_5	21 turns close wound tapped 13 t., 7 t.	18 turns spaced to 1½ in. tapped 6 t., 5½ t.	9 turns spaced to 1½ in. tapped 3 t., 2½ t.	3 turns spaced to 1½ in. tapped 1½ t., 1 t.
C_4	35 μμF.	35 μμF.	10 μμF.	10 μμF.
C_5	35 μμF.	35 μμF.	10 μμF.	10 μμF.
C_6	75 μμF.	75 μμF.	75 μμF.	50 μμF.

CHAPTER 8 POWER SUPPLIES

SMALL receivers may be operated direct from dry batteries, both for L.T. and for H.T., but if the consumption is so heavy that dry batteries are uneconomical accumulators may be used instead. The use of accumulators for H.T. supplies is rather troublesome, and a vibrator or a motor-generator is usually preferred. Public supply mains are, of course, the best sources of power, but careful thought must be given to the question of earthing where D.C. mains are used.

Dry Batteries

The 1·4-volt series of valves are designed to operate from dry batteries: the total filament current drain in a simple receiver is usually not more than 0·1–0·2 A. A dry battery H.T. supply, giving a total current of up to 20 mA. at 90–120 volts, is satisfactory for all such receivers, and often the consumption can be kept below 10 mA.

Since the voltage of a dry battery falls appreciably throughout its life, it is important to check the voltages at frequent intervals. The performance of a receiver will deteriorate as the voltage falls, but the change will perhaps not be noticed, for the simple reason that it occurs very slowly. A filament rheostat may be included in the circuit to permit the filament current to be adjusted to the minimum practicable value at all times.

It is worth remembering that dry batteries last much longer if they are used only for short periods with long intervals of rest rather than for long unbroken periods of operation. Moreover, it is false economy to use undersized batteries, especially for the filament supply.

Accumulators

The 2-volt series of valves are designed to operate on an accumulator supply. The popular 6·3-volt range, although mostly used in mains-operated receivers, was originally designed to run on 6-volt car batteries and is, of course, very convenient for car receivers. The nominal heater voltage was fixed at 6·3 volts because this is the average voltage of a 6-volt car battery while in normal use, *i.e.* on charge.

Accumulators are suitable for a current drain of up to about 10 A. For larger currents, the battery would be inconveniently heavy or the discharge time would be too short. The largest size of battery practicable for amateur purposes is of 100–120 amp-hour rating.

The use of H.T. accumulators is clumsy and expensive.

Rotary Converters

The rotary converter is a convenient means of deriving an H.T. supply from either an accumulator or D.C. mains. The efficiency is roughly 60 per cent., and if the valve heaters are also to be supplied from the output of the converter (as, for example, where 13-volt heaters have to be fed from a

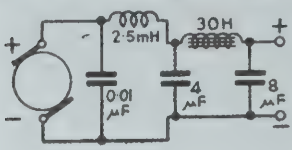


Fig. 86.

The output circuit of a rotary converter, showing a "hash" filter consisting of a 2·5 mH. R.F. choke and a 0·01 μF. condenser. A smoothing circuit is included to remove the commutator hum.

6-volt battery), the total amount of power may be about 50 watts and quite a large battery would then be necessary.

Small rotary converters are eminently suitable for running "car radio" receivers.

Although the ripple frequency of the D.C. output is relatively high (*e.g.* 500 c/s.), it is often necessary to use an elaborate filter to eliminate the troublesome "hash." A typical filter circuit for this purpose is shown in Fig. 86.

Vibrators

As an alternative to the rotary converter for deriving H.T. from a low-voltage battery, a vibrator in combination with a transformer and rectifier may be used. The vibrator serves to interrupt the D.C. from the battery at a suitably high rate (usually about 110 c/s.). This interrupted D.C. is passed through the primary of a step-up transformer where it behaves like A.C., and the output from the secondary is A.C. of similar frequency but of higher voltage. The waveform is not sinusoidal and is almost rectangular: numerous strong harmonics are therefore present.

Rectification of the output current may be effected by the usual type of power rectifier valve or by a set of contacts carried by the vibrating armature: see Fig. 87.

The primary winding of the transformer is centre-tapped so that the direction of the battery current can be reversed; the vibrating reed closes the circuit at each end of its swing, the current passing through each half of the primary in turn. The reed is energised by a high-resistance electro-magnet connected to the battery. When the current is first switched on, the electromagnet deflects the reed until the contact is reached: this short-circuits the magnet winding and the reed swings back and is carried by its momentum to the contact on the opposite side, thus causing the electro-magnet to be energised again. The speed of vibration and therefore the frequency of alternation of the current through the transformer depends on the mechanical resonance of the reed.

Fig. 87B shows the so-called synchronous or self-rectifying vibrator arrangement. Here the secondary current is rectified by the intermittent closing of the circuit by the second pair of contacts. Since the vibrating element is identical with that which controls the primary current, the matching of the periodicities is automatic.

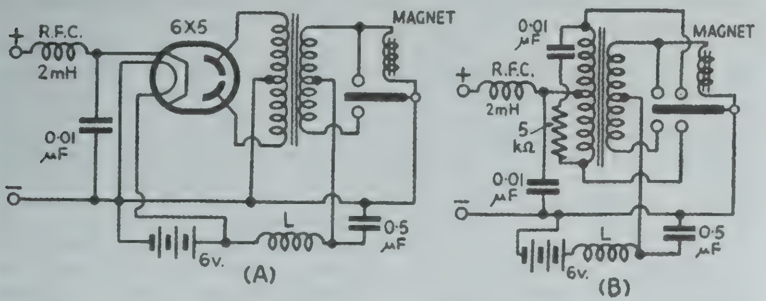


Fig. 87.
Vibrator power supply circuits. The output is passed through a conventional type of smoothing circuit to remove the low-frequency ripple. The R.F. choke L may consist of 40 turns of No. 16 S.W.G. D.C.C. copper wire close-wound on a 1-in. former.

An R.F. choke (wound with thick wire so as to carry the heavy current) and a by-pass condenser are connected in the primary circuit in order to suppress the " hash " generated by the slight arcing at the contacts. This arcing is itself minimised by a condenser and a resistance shunted across the contacts.

An ordinary type of mains transformer may be used with a vibrator if it has a spare winding available for the battery input, a 6·3-volt winding being suitable for a 6-volt battery. In this way the power supply for a receiver can be designed to operate on A.C. mains or a battery with only the simplest of switching arrangements.

Vibrator supplies are capable of providing the H.T. current for any receiver of moderate power consumption, but the voltage regulation is not very good owing to the non-sinusoidal waveform. The current load should therefore be made as light as possible, and this will also lengthen the life of the contacts.

D.C. Mains

For most receivers the voltage of the usual D.C. mains supply (200 volts) is adequate. A " hash " filter is often necessary to suppress the interference originating at the generating station and carried by the line: see Fig. 88.

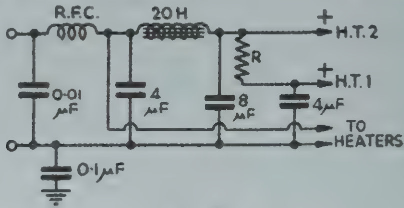


Fig. 88.
A power supply circuit suitable for D.C. mains. The R.F. choke should be capable of carrying the heater current without becoming over-heated. The resistance R must be chosen to suit the current load at the reduced voltage H.T.1.

When the positive side of the supply is earthed special care must be taken, for all the components in the receiver which are usually at earth potential in A.C. mains receivers—including the chassis—will be at a high potential to earth. A large condenser (about 0·005 μF.) should be connected between the chassis and earth to serve as a by-pass for R.F. currents.

If the valve heaters are to be fed from the D.C. mains, it is obviously worth while to connect them in series rather than in parallel. This is quite practicable with the 6·3-volt range but it is preferable to use the special types designed for this purpose: these special types are rated at 12–13 volts and 25–26 volts, the current being 0·15 A. or 0·3 A.

It is essential to see that each heater has the proper current passing through it. Thus it may be necessary to connect some of the heaters in series and place a resistance in parallel with them so as to match the current passing through another valve which has a heavier current rating.

A series resistance must be added to the chain of valve heaters to accommodate the remaining voltage-drop. The resistance may very conveniently be constituted by a barretter which will offer the further advantage of stabilising the current against variations in the mains voltage. Such a barretter must be chosen carefully to match the current load and the expected range of voltage-drop.

Alternatively, or in addition, a fixed resistance can be used: this may be in the form of the so-called *line cord*, which is a flexible twin-lead made of resistance wire instead of copper. Its length must be accurately measured so as to introduce the required amount of resistance into the circuit. A considerable quantity of heat has to be dissipated by the line cord and it must therefore not be coiled up or kept in a confined space when in use.

A.C./D.C. Supply Circuits

It is sometimes an advantage to be able to operate a receiver on either A.C. or D.C. mains without alteration to the internal connections. If the valve heaters are arranged in series (or possibly partly in series-parallel) they may be energised equally well by A.C. or D.C., but obviously a rectifier will be required for the H.T. supply for A.C. operation although it is not necessary for D.C. operation. In order to make the receiver truly universal,

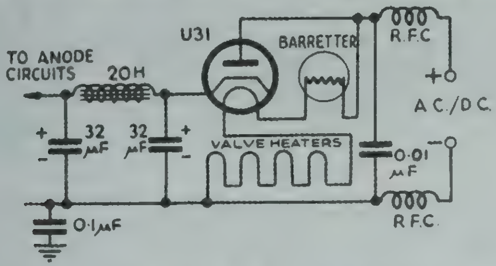


Fig. 89.
The usual form of A.C./D.C. supply circuit, showing the series connection of the valve heaters. The barretter may be replaced by a fixed resistance the value of which must be selected to give the required heater current.

it is generally considered worth while to leave the rectifier connected in circuit when a D.C. supply is used: the voltage-drop across the rectifier is only 20–30 volts. Moreover, the inclusion of the rectifier sometimes serves a useful purpose in removing the high-pitched hum which is otherwise audible due to the commutator ripple on the D.C. supply voltage.

A circuit arrangement which has been found satisfactory in universal receivers is shown in Fig. 89.

A.C. Mains

When A.C. mains are used for the power supply in a receiver, it is essential that the valves should have indirectly-heated cathodes: this avoids the introduction of " mains hum " into the circuit. By way of exception, the output stage in an A.F. amplifier may have a filament-type cathode: any hum introduced here from the A.C. flowing through the filament is not likely to be very great (since the gain in this stage is usually small), and any hum that may be present can be eliminated by earthing the electrical centre of the heater supply, as shown in Fig. 90.

In all other amplifier stages, except the output stage, and in detectors, oscillators, etc., it is usually satisfactory to earth one side of the heater supply. A suitable by-pass capacitance is connected across the heater to avoid trouble due to stray R.F. currents: see Fig. 91. R.F. chokes are not likely to be needed in the heater circuit wiring in any receiver which is restricted to frequencies below about 40 Mc/s.

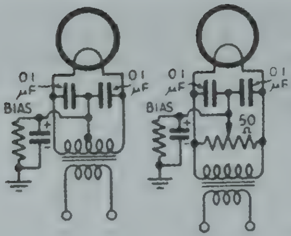


Fig. 90.

To remove hum arising from the use of A.C. for filament heating in an A.F. output valve, it is usual to make the cathode connection to the centre of the filament winding, or better, to a potentiometer connected across the filament.

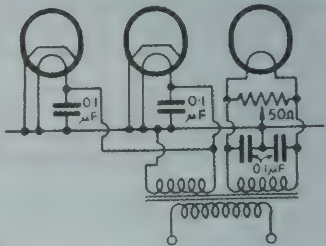


Fig. 91.

Indirectly-heated valves may have their heaters shunted by a by-pass condenser, one side being connected to earth. A directly-heated output valve with a balanced cathode connection should then be fed from a separate secondary winding.

For the H.T. supply a full-wave rectifier (either a valve or a metal rectifier) is generally used, although for monitors and E.H.T. supplies in cathode-ray oscillographs and television receivers, where the current drain is small, a half-wave rectifier is quite satisfactory. Typical filter circuits for smoothing out the ripple are shown in Fig. 92.

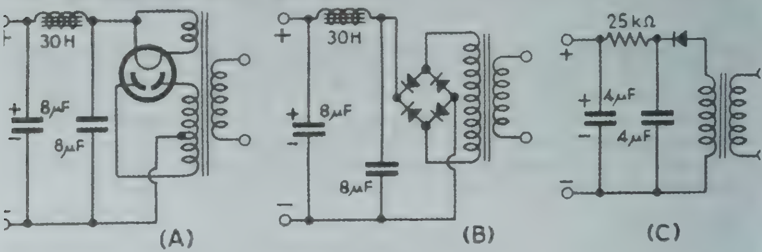


Fig. 92.

H.T. supply circuits for A.C. mains receivers. The most popular arrangement, shown at (A), uses a bi-phase valve rectifier. The four-arm metal-rectifier arrangement shown at (B) is an efficient substitute. For very small currents, a half-wave rectifier and a resistance-capacitance filter (C) are often used.

Very roughly, the output voltage obtainable from a circuit of the kind shown in Fig. 92A can be assumed to be the same as the nominal R.M.S. voltage of each half of the transformer secondary. Precise calculation is not easy, for the output voltage falls as the current is increased. By using a choke-input type of filter, this dependence of output voltage upon the current load is appreciably diminished; *i.e.* the *regulation* is improved. The choke-input filter, however, results in a lower output voltage than the condenser-input filters shown in Fig. 92.

While the choke-input filter is very much to be preferred in high-power A.F. amplifiers (especially Class B), the improvement in voltage stability is not of great benefit in the design of stable R.F. oscillators. For this purpose, a special effort must be made to keep the voltage constant, irrespective of variation in the output current or in the mains voltage.

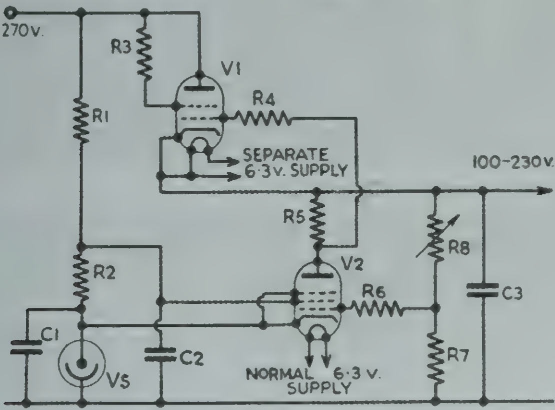


Fig. 93.

A voltage-regulator circuit suitable for controlling the output from a conventional power supply unit. The voltage of the supply to the receiver can be adjusted between 100 and 230 volts approx. for an input voltage of 270 volts. The current load should not exceed 70 mA. (*Note.*—The connections to the stabiliser tube should be reversed.)

R1, 2	25,000 ohms 1-watt.	V1	6L6 or EL37	C1	0·01 μF. mica.
R3	100 ohms 1-watt.	V2	EF50.	C2	4 μF.
R4, 6, 7	30,000 ohms (R7 1-watt).			C3	2 μF.
R8	50,000 ohms wire-wound variable.			Vs	7475 or 85A1 *Mullard*

Voltage Stabilisers

The usual type of voltage stabiliser depends for its action on a variable series resistance, the value of which is automatically controlled by the input voltage or the load current. Fig. 93 shows a typical voltage stabiliser circuit suitable for feeding the local oscillator in a superhet, a frequency meter, etc.

A less effective but still useful degree of stabilisation can be obtained by the use of a simple neon-filled regulator tube: see Fig. 94. The output voltage is restricted to within rather narrow limits set by the characteristics of the tube. Higher voltages than the rating of a single tube can be obtained by connecting two or more tubes in series.

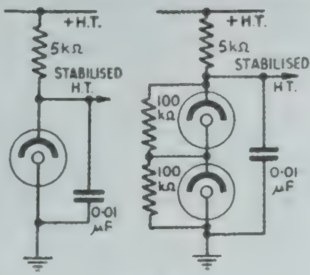


Fig. 94.

Simple neon-tube voltage-stabiliser circuits.

CHAPTER 9 FAULT-FINDING, ADJUSTMENTS AND CALIBRATION

FAULTS may be divided into three classes: (i) the insidious kind that causes a gradual decline in receiver performance which is so gradual that it escapes notice, (ii) the sudden fault which is so readily apparent that it is impossible to ignore it, and (iii) the intermittent fault which may occur at any moment yet obstinately refuses to appear whenever an attempt is made to locate it. This elusive, intermittent kind of fault is unquestionably the most troublesome, and its removal can be effected with certainty only by a painstaking and thorough checking of all the suspected components.

General Testing and Fault-finding

Unless it is known with reasonable assurance where a circuit needs attention, the best procedure is to begin at the output and work back to the beginning.

The headphones or loudspeaker can be tested with a small dry battery or preferably A.C. (from an A.F. oscillator or the filament winding of a transformer). The output valve should be tested, first in its anode circuit (with a suitable milliammeter) and then in its grid circuit by applying a small A.F. voltage.

Next the intervalve coupling by which the output valve is driven from the preceding valve should be examined. This may mean the testing of a condenser, a resistance or a transformer. The testing of certain components may present a rather difficult problem if suitable equipment is not available (*e.g.* a capacitance or impedance bridge), and it is often better to try a temporary substitution by a component which is known to be in sound condition.

When the circuit has been checked as far back as the detector, it can be tested by applying a modulated R.F. signal to the detector grid (remembering, of course, that the control grid should never be left " floating " but should at least be connected to the cathode or to earth through a resistance of a megohm or two). If a modulated oscillator is not available, a simple resonant circuit tuned to the local broadcasting station will provide a useful supply of modulated R.F. test signals.

The detector tuning circuit can be tested by loosely coupling the aerial to it, using any incoming signal as a test signal, but when the frequency is

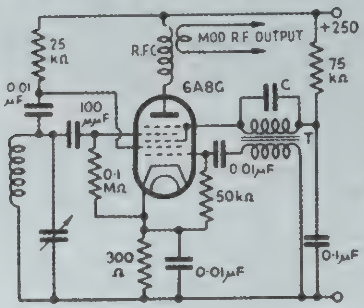


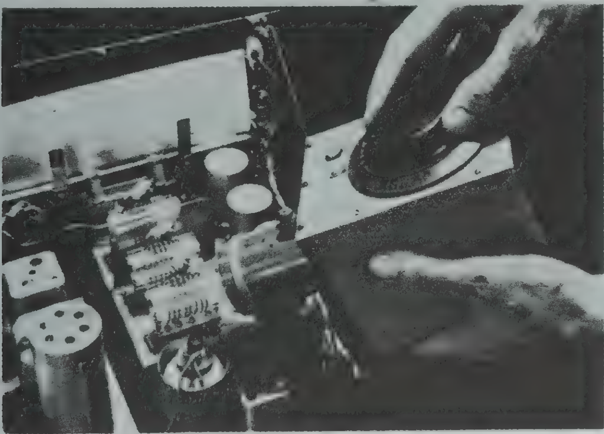
Fig. 95.

A modulated test oscillator using the transitron circuit. A.F. oscillations are produced by feed-back in the intervalve-type transformer T, tuned by a condenser C of suitable value. The R.F. output transformer may be made from an " all-wave " choke, one section of it being disconnected and made to serve as a low-impedance winding.

more than 2–3 Mc/s. this method is not always satisfactory because the signal strength will probably vary quite markedly due to fading.

A suitable circuit arrangement for a modulated oscillator is shown in Fig. 95. By fitting different inductances, and perhaps also changing to a different variable condenser, the same oscillator can be made to cover all the frequency ranges that are likely to be needed (say 50 kc/s. to 15 Mc/s.).

For testing the I.F. amplifier in a superhet, such a modulated oscillator is almost essential. The same principle of working backwards to the beginning should be followed, especially when the I.F. amplifier comprises two or more stages, the frequency of the oscillator being set, of course, at the required peak frequency of the I.F. amplifier. If a crystal filter is to be used, it is a good plan to remove the crystal temporarily and connect it into the test oscillator so that the I.F. amplifier can be set accurately to the crystal frequency.



Measuring the oscillator frequency with an absorption type frequency meter.

When testing the I.F. amplifier, the mixer valve should be regarded as its first stage: *i.e.* a modulated signal at the I.F. amplifier frequency should be applied to the control grid of the mixer (after having disconnected the signal-frequency circuit).

The signal-frequency circuits may be checked with a modulated test oscillator, the sequence of testing being continued backwards until the aerial circuit is reached.

Faults in Valves

All valves, except low-wattage battery valves, become appreciably heated in use due to the power dissipated by the filament or the cathode heater. If a valve is warm to the touch, it can therefore be assumed at least that the cathode is being normally heated. The anode current may be sufficient—and it very often is—to raise the temperature of the bulb so far as to make it unpleasantly hot to the touch: the extra heat is derived from the power dissipated by the anode. This is obviously another useful guide. An output valve that does not become quite hot to the touch may be suffering from loss of cathode emission or some fault may have developed in the circuit.

Intermittent internal contacts or loosely mounted electrodes may be detected by tapping the bulb. The effect of such faults on the performance of a receiver is usually apparent in a very marked degree.

Faults in R.F. Amplifiers

Probably most of the troubles encountered in R.F. amplifiers are due to unintentional feedback. This causes self-oscillation in bad cases, and may result in erratic behaviour of the tuning circuits, or may lead to peculiarities in general behaviour due to unexpectedly high selectivity. If self-oscillation occurs in I.F. amplifiers, all C.W. carriers will be audible even with the B.F.O. switched off, and spurious carriers may appear due to the heterodyning of the local oscillator by harmonics of the oscillating I.F. amplifier.

Such faults may arise from ineffective R.F. by-passing of the screen-feed or anode-feed circuits, or to coupling between the grid and anode circuits of any one stage, or to coupling between different stages. All earth-return leads must be as short as possible and all joints must be well made so as to have very low resistance.

Unintentional feedback may be occurring in two or more places simultaneously, and to simplify the task of locating the trouble it is worth while to earth all control grids which are not serving any useful purpose while the test is being made. Alternatively any unnecessary valves can be removed from their sockets (provided that their removal does not normally affect the operating conditions of the valve that is being tested).

The absorption frequency meter is particularly useful in locating unwanted oscillation: when it is coupled to the offending circuit and tuned through the frequency of the spurious oscillation, the fault will temporarily disappear, or at least show some change such as a slight shift in the heterodyne frequency.

Sometimes the self-oscillation may be curable only by the insertion of a resistance in the control-grid lead or the anode lead. Such a resistance is called a *stopper*, and may be about 10,000 ohms if in a control-grid lead or about 20–100 ohms if in an anode lead. Stoppers are not advisable in the lead to a screen grid. The oscillation which is amenable to this form of treatment is usually at very high frequency, and as an alternative to the resistance it may be preferable to use a miniature V.H.F. choke. In any case, the stopper resistance or choke should be placed as near as possible to the valve pin.

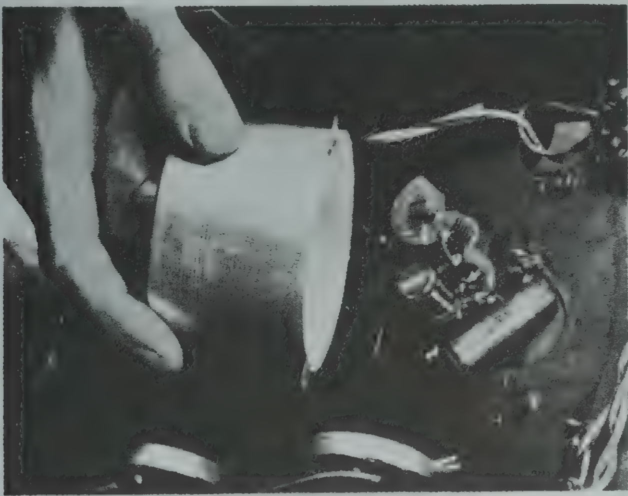
If it seems difficult to prevent feedback entirely, it may help to reverse the connections to some of the by-pass condensers, especially if they are of the tubular form, since it is advantageous to have the outside foil electrode at earth potential whereby it acts as a screen for the high-potential electrode.

Faults in A.F. Amplifiers

Similar precautions should be taken in A.F. amplifiers to those in R.F. amplifiers, but there is considerably less risk of trouble by reason of the much greater impedance of the stray capacitances and the negligible impedance of the connecting wires at audio frequencies.

Self-oscillation may occur in high-gain A.F. amplifiers if positive feedback is present. This will usually take the form of an audible howl or whistle (independent of tuning adjustments in the R.F. part of the receiver). If only a small amount of positive feedback is present, the effect may be merely the accentuation of the higher audio frequencies. The fault may be cured by ensuring that all the screen and anode feed circuits are adequately by-passed

to earth, especial care being necessary to provide thorough de-coupling if the amplifier has two or more stages. Electromagnetic coupling between two A.F. transformers may be responsible for the unwanted feedback, and the remedy in this case is the re-arrangement of the transformers so as to minimise the coupling. This is best done experimentally, observing the performance of the receiver as each transformer is set at various angles and moved to various positions.



A coupling transformer in the anode circuit of the first frequency changer in a double superhet It is tuned to 5 Mc/s. and a 70-ohm line passes the signal to the first I.F. amplifier.

It is possible that an A.F. amplifier may be generating super-audible oscillations. The effect will be apparent as a form of distortion, and the existence of such oscillations may be detected by connecting a milliammeter in the anode circuit and observing whether there is any change in the anode current when the grid terminal is touched with the finger.

Faults in Detectors

Perhaps the most common trouble experienced in detectors is the appearance of A.C. mains hum. Leakage between the heater and the cathode in an A.C. type of valve is one of the usual forms of trouble: the hum may be induced by resistance leakage or by electromagnetic induction. These faults may be curable by adequate by-passing of the heater with a suitably large capacitance ($0 \cdot 001$–$0 \cdot 01$ μF.), or by reversing the connections to the heater. One side of the heater circuit is usually connected to the chassis and this connection is best made in the proximity of the detector circuit. If all attempts to remove the hum by modification of the circuit should fail, the only alternative is to try another valve.

Some valves are prone to microphonic effects. A felt or rubber cover closely fitting over the bulb may provide a complete cure.

In regenerative detectors, as used in straight receivers, the feedback control may exhibit "overlap." A change of grid leak, or the introduction of a small standing bias (*e.g.* by a cathode resistance suitably by-passed with a condenser), or a reduction of the impedance in the anode circuit, will usually remove this fault.

Faults in Tuning Circuits

In band-switched receivers, the tuning control may behave in a peculiar way if there is appreciable coupling between an unused coil and the coil which is actually switched into circuit by reason of the stray capacitance between

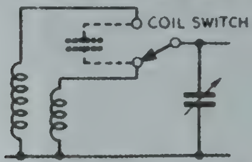


Fig. 96.

Inductances not actually in use may be coupled sufficiently closely to the selected coil by the small capacitance between the switch contacts to produce spurious resonances.

the switch contacts and the associated wiring: see Fig. 96. If this fault occurs in an oscillator (as in a superhet or a regenerative detector in a straight receiver), it should be easily apparent by the nature of the frequency variation throughout the range of the tuning condenser. In an R.F. amplifier, however, this unwanted coupling may be much less obvious but equally objectionable since it will cause a loss of sensitivity over some part of the tuning range.

Faults of this kind can be cured most effectively by providing a further switching arrangement for short-circuiting the coils which are not actually in use. It should be sufficient to provide a short-circuit only for the coil which is the next largest in inductance value to the one in use.

Faults in Switches

Rotary selector switches become dirty after long use and produce scratching noises in the circuit when moved, sometimes causing erratic variations of frequency if the fault occurs in an oscillator or detector circuit. The dirt can readily be removed by washing the contacts with carbon tetrachloride applied with a small camel-hair brush.

Faults in Power Supplies

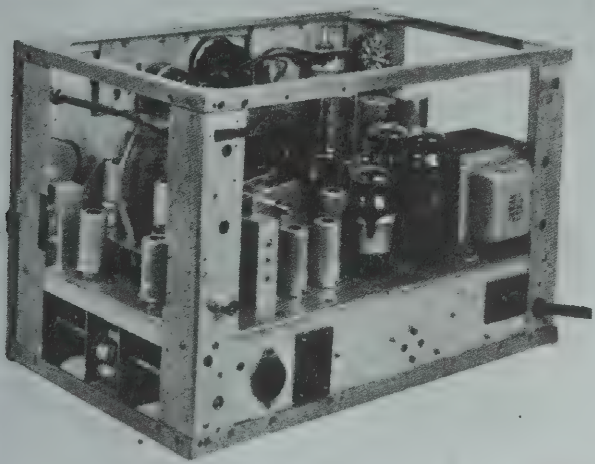
The presence of a strong A.C. mains hum in the headphones or loudspeaker does not necessarily indicate a fault in the power supply: it may arise from faulty heater-cathode insulation in one of the valves or from defective decoupling. To locate the cause of A.C. hum it may be quickest—in the absence of any suspicious behaviour in the receiver itself—to examine the power unit first.

Electrolytic condensers are perhaps the most likely source of trouble. The simplest procedure is to replace the suspected condenser by another of similar rating. If the hum is still present, the smoothing choke (or resistance) should be tested. Failing an inductance meter or an impedance bridge, a simple test can be made by short-circuiting the choke while in use. If the hum does not markedly increase, it is fairly certain that the choke has developed an internal short-circuit.

A further possible cause of A.C. mains hum is the increased load on the filter which would result from a short-circuit of the H.T. feed. The smoothing

effect of the filter diminishes as the current drain increases: in a well-designed system, however, the increase in hum should not be very great even for a considerable increase in the current load.

The cathode emission in a rectifier valve may diminish with age to a point at which the resultant increase in the valve impedance causes a lowering of the output power from the supply unit. This is best checked by measuring



(Courtesy *Denco Ltd.*)

A rear view of the receiver illustrated on p 67. The power supply components, seen at the extreme right, are so placed that they can be adequately cooled and are well removed from the oscillator in order to keep frequency drift at a minimum.

the feed current and the voltage, and if the values have fallen the only reliable remedy is to fit a new valve. Care should be taken, of course, to ensure that the mains voltage is normal while the measurements are being made.

Faults in Miscellaneous Components

Rotary potentiometers used as gain controls sometimes become " noisy " after long use. This is due to dirt on the sliding contact and the surface of the resistance element, and may be cured by washing with carbon tetrachloride. A very high resistance potentiometer is rather more liable to give trouble if the sliding contact is made to carry D.C. than if it is made to carry only A.C. (*i.e.* either R.F. or A.F. currents).

Crackling noises in the headphones may be due to a fault in the windings or in an A.F. transformer. A makeshift cure can sometimes be effected by short-circuiting the faulty winding in the headphones—since there are usually two windings in series in each earpiece—but the best procedure is to have the faulty bobbins rewound or to obtain a replacement.

Some valve sockets are liable to give trouble by the spring elements failing to bear properly on the valve pins: careful prodding with a rod of insulating material should disclose faults of this kind.

Frequency Measurement

In the adjustment and calibration of a receiver it may be necessary to measure the frequency of a self-generating circuit (*e.g.* the local oscillator in a superhet) or the frequency of a resonant circuit which does not oscillate on its own account (*e.g.* a grid tuning circuit in a straight receiver).

Three types of frequency meter are commonly used for these purposes—

(i) Grid-dip oscillator—for non-generating circuits.
(ii) Heterodyne frequency meter—for self-oscillating circuits.
(iii) Absorption frequency meter—for self-oscillating circuits.

The grid-dip oscillator operates by virtue of the fact that a resonant circuit coupled to it will absorb energy from it when the frequency of the oscillator coincides with the resonant frequency. The most sensitive indicator of the fact that energy is being absorbed is the rectified grid current: see Fig. 97. Generally, inductive coupling is likely to be convenient but

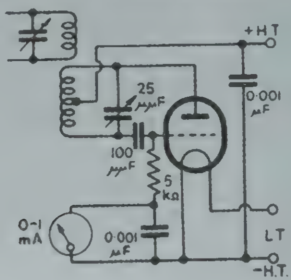


Fig. 97.
A grid-dip oscillator circuit. Any miniature triode (or pentode connected as a triode) may be used. For A.C. heating, an indirectly-heated type is preferable.

Coil dimensions:

3·5 Mc/s.	...	...	60 turns No. 28 S.W.G.
7 Mc/s.	...	...	34 turns No. 24 S.W.G.
14 Mc/s.	...	...	14 turns No. 24 S.W.G.
28 Mc/s.	...	...	6 turns No. 24 S.W.G.

All coils are centre-tapped and are close-wound on cylindrical formers of 1½ in. diameter.

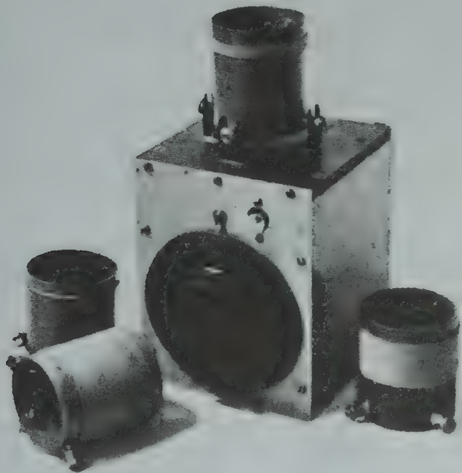
capacitive coupling is also permissible (using the smallest capacitance necessary to give a reliable indication so as to avoid a possible detuning of the circuit).

The aligning of the signal-frequency circuits of a superhet is made much easier by the use of the grid-dip oscillator, and the need for a signal generator for this purpose practically disappears.

To measure the frequency of a circuit which is itself generating oscillations, a heterodyne frequency meter is necessary if the measurement must be made with a high order of accuracy, but if the tolerable error is greater than about 0·5 per cent. it may be satisfactory to use an absorption frequency meter.

The heterodyne type differs very little from the calibrated grid-dip oscillator, the only essential difference being the provision of headphones. It is, in effect, a calibrated C.W. receiver of relatively low sensitivity but with a high order of frequency stability. The signal to be measured is found in the ordinary way by searching for the heterodyne beat note. The setting which gives the central zero beat corresponds to the exact frequency of the signal.

Care must be taken to verify that it is the fundamental frequency of the signal that is being heard and not a harmonic frequency. A check measurement by an absorption frequency meter will remove any doubt, for this type of meter is ineffective at harmonic frequencies.



An absorption type frequency meter. The variable condenser has a capacitance range of 12–500 $\mu\mu$F., and six interchangeable coils are used for the complete range from 90 kc/s. to 65 Mc/s. The serviceable range covered by each coil corresponds to a frequency ratio of approximately 4 : 1.

The absorption frequency meter consists in its simplest form of a calibrated tuned circuit, usually a fixed inductance and a variable condenser. If this is coupled to an oscillating circuit, there will be an absorption of energy at the resonant frequency and a consequent change in the running conditions of the oscillator, such as the grid current or the anode current, which can be used as an indication of resonance.

Alternatively, the absorption frequency meter may be provided with a rectifier and a microammeter for indicating the setting at which the maximum energy is absorbed: see Fig. 98.

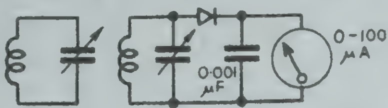


Fig. 98.
The simplest form of absorption frequency meter, shown on the left, relies on observations in the circuit to which it is coupled to indicate resonance. The more elaborate circuit shown on the right contains its own indicator.

Calibration of Tuning Scales

In the alignment and calibration of a superhet it is almost essential to have either a heterodyne frequency meter or an absorption frequency meter for measuring the frequency of the local oscillator. It should be remembered, of course, that the oscillator frequency will be higher or lower than the signal frequency, according to the design, by an amount equal to the peak frequency of the I.F. amplifier.

In the calibration of a straight receiver, it is merely necessary to calibrate the oscillator (which is usually the detector itself) and to ensure that all the other tuned circuits are correctly tracked (*i.e.* tuned to resonate at the same frequency as the detector).

It is worth while aiming at the highest accuracy possible in the calibration. If a suitable tuning scale is fitted, and if the frequency stability is sufficiently high, it should be possible to mark the scale to show the frequency correct to within ± 5 kc/s. or even less. This is a valuable feature when searching for a station of known frequency.

Ganged Tuning

The alignment of a straight receiver is a relatively simple matter. Assuming that the circuits are simultaneously tuned by a ganged condenser, it is desirable to be able to vary the frequency of any one of the circuits above and below the value as determined by a particular setting of the tuning condenser but without altering that setting. This can easily be done by temporarily inserting a brass or copper plug into the field of the inductance to raise the frequency and an iron-dust plug to lower the frequency: alternatively, to lower the frequency, a miniature variable condenser may be connected across the coil terminals. If a stronger signal is obtainable with the aid of the plug or the shunted capacitance, it is evident that the circuit is not correctly aligned, and the remedy should be obvious.

If correct alignment cannot be achieved merely by adjustment of the trimmers, the inductance values may be at fault. A series of check measurements may be made to verify the resonant frequencies of the various circuits by means of a grid-dip oscillator.

The procedure for checking the ganged tuning in a superhet is considerably more complicated. The first step is to see that the oscillator covers the necessary tuning range. Where the receiver is designed for wide frequency coverage, the oscillator circuit will contain a padding condenser. This should be adjusted to set the minimum frequency (when the tuning condenser is at its maximum capacitance), and the trimming condenser should be adjusted to set the maximum frequency (when the tuning condenser is at its minimum capacitance).

The signal frequency circuits are next adjusted by means of their trimming condensers at the maximum frequency. If the tracking is appreciably in error at the minimum frequency, it will probably be necessary to alter the value of the signal frequency inductances. In this case the trimmers will, of course, have to be readjusted when the signal frequency inductances bear the correct ratio to the oscillator inductance. The tracking should then be reasonably good over the entire tuning range, minor adjustments being confined to the trimming condensers and the padding condenser. In order to maintain the frequency calibration it is obviously very desirable not to make any further adjustments to the oscillator circuit.

When the receiver is restricted to the narrow ranges of the amateur bands, there will be no padding condenser. In making the adjustments for the

proper oscillator tuning range, special attention must be given to the desired amount of band-spread. Correct tracking of the signal frequency circuits is then obtained by adjustment of their trimmers and their inductance values.

The procedure in all these ganged-tuning operations may prove to be laborious and it is wise not to begin unless ample time is available. An orderly approach and some patience are helpful if not absolutely essential, for the performance of a receiver can quickly be marred by ill-considered tinkering with the tracking elements. If the tuning circuits should become completely out of alignment, it will be simpler to work through the adjustments from the beginning, preferably with the aid of an accurately calibrated grid-dip oscillator.

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Interpolation made easy. A. G. Dunn. March, 1949, p. 222.
R.F. pentode valves. H. N. Gant. March, 1949, p. 219.
Frequency modulation discriminator. B. W. St. Leger Montague. March, 1949, p. 217.
True S-metering. K. E. Marcus. February, 1949, p. 198.
A narrow band F.M. adaptor. E. W. Nield. December, 1948, p. 138.
A communication superhet from the TR1196. R. M. Baerlein. November, 1948, p. 113.
An all-band aerial for the B.R.S. G. A. Hook. October, 1948, p. 91.
A grid-dip oscillator. G. P. Anderson. September, 1948, p. 63.
Absorption meters. C. H. L. Edwards. July, 1948, p. 15.
A portable communications type receiver. A. M.-M. Payne. July, 1948, p. 5.
The whys and wherefores of "S" meters. J. N. Walker. June, 1948, p. 234.
An amateur-bands straight receiver. A. Alsbury. April, 1948, p. 196.
The *Radiovision* "Hambander." J. C. February, 1948, p. 158.
A receiver for the amateur bands. L. F. Worssam. January, 1948, p. 130.
The B.C.348 communication receiver. G. L. Benbow. November, 1947, p. 91.
The AR88 noise limiter circuit. E. W. Nield. September, 1947, p. 46.
The B2 transmitter-receiver. C. H. L. Edwards. September, 1947, p. 44.
A variable selectivity I.F. amplifier using modified R1155 transformers. A. J. Bayliss. July, 1947, p. 4.
The *Eddystone* S640 receiver. June, 1947, p. 208.
Plug-in coils for home-built superhets. G. C. Eyre. May, 1947, p. 183.
Coil inductance measurements. P. Newton Nield. May, 1947, p. 183.
A portable direction finding receiver. J. M. S. Watson. April, 1947, p. 161.
Noise signal generator. W. P. Dolphin. April, 1947, p. 158.
An R-C audio frequency filter. P. Newton Nield. March, 1947, p. 144.
The anode follower. B. H. Briggs. March, 1947, p. 138.
The R1155 receiver modified for amateur use. A. H. Mason. November, 1946, p. 70.

A 3-valve superhet. B. K. Sands. October, 1946, p. 59.

An amateur-built eight-valve communications superhet. E. W. Nield. October, 1946, p. 50.

Station design and planning—IV: The amateur bands receiver. W. H. Allen. March, 1946, p. 138.

A low distortion diode detector. R. A. Knowles. January, 1946, p. 108.

Radio receiver specifications. A. Stuart McNicol. November, 1945, p. 66.

Communications receivers. R. H. Hammans. August, 1945, p. 18.

The LC circuit: frequency coverage. R. H. Hammans. November, 1944, p. 71.

A compact broadcast receiver. A. G. Dunn. September, 1944, p. 42.

Eddystone communications receivers. R. Heatley. August, 1944, p. 24.

Diode detection and automatic volume control. W. G. Johnson. July, 1944, p. 11.

Servicing communication receivers. H. W. Miles. July, 1944, p. 7; August, 1944, p. 23.

Negative feedback in transmitters and receivers. H. A. M. Clark. July, 1944, p. 2; August, 1944, p. 18; September, 1944, p. 34.

A four-valve communications receiver. W. G. Johnson. January, 1944, p. 98.

The Synthescope. R. H. Hammans. December, 1943, p. 82.

Theme and variations on the D.S.H. I. B. Whitstable. September, 1943, p. 39.

Applied D.C. F. J. Forbes. September, 1943, p. 37; November, 1943, p. 70; December, 1943, p. 88.

A simple noise limiter. J. W. Cropper. September, 1943, p. 36.

The principles of " single-span " tuning applied to the higher frequencies. 2FPP. July, 1943, p. 5.

An austerity pre-selector. R. G. Kitchen. May, 1943, p. 167.

Some " Heterofil " snags and their solution. F. Dearlove. May, 1943, p. 164.

Trouble tracing. A. H. Mason. November, 1942, p. 70; December, 1942, p. 86; January, 1943, p. 98.

An efficient one-valve short-wave detector unit. S. E. Janes. December, 1941, p. 200.

A 4-valve band-switching receiver for A.C. mains. S. K. Lewer. November, 1941, p. 160.

The " Heterofil." R. L. Varney. June, 1941, p. 394.

An efficient receiving installation. A. W. Birt. May, 1941, p. 354.

Two-stage preselector with band switching. T. B. Wimbush. January, 1941, p. 213.

Construction of a crystal bandpass filter. P. F. Cundy. November, 1940, p. 134.

Compact two-valve receiver. W. E. Bartholomew. September, 1940, p. 73.

A.V.C. with plate and grid detectors. S. A. G. Cook. June, 1940, p. 462.

Receiver stability. J. N. Walker. February, 1940, p. 331; March, 1940, p. 368.

Crystal bandpass filters. E. L. Gardiner. August, 1939, p. 75; September, 1939, p. 141; October, 1939, p. 178; November, 1939, p. 213; December, 1939, p. 251.

Quiet beat-frequency oscillator circuit. E. L. Gardiner. December, 1939, p. 246.

PROCEEDINGS OF THE R.S.G.B.

Practical use of frequency modulation on amateur frequencies. D. N. Corfield. Autumn, 1948, p. 1.

QST

A simplified circuit for audio image rejection. G. Grammer. September, 1949, p. 13.
Audio filters for eliminating QRM. R. R. Bennett. July, 1949, p. 51.
A fixed-tuned plug-in converter. J. Aletto. July, 1949, p. 62.
Modernising the pre-war HRO. L. G. Windom. June, 1949, p. 51.
A compact converter for 6 and 10. C. V. Chambers. February, 1949, p. 23.
" Souping up " a war-surplus HRO. P. D. Rockwell. February, 1949, p. 39.
New life for old receivers. B. Goodman. December, 1948, p. 16.
A peaked audio amplifier for communication receivers. G. D. Hanchett. September, 1948, p. 16.
Triple conversion for the communications receiver. W. I. Orr. September, 1948, p. 53.
A super-selective C.W. receiver. T. A. Githens. August, 1948, p. 16.
Practical single-sideband reception. D. E. Norgaard. July, 1948, p. 11.
A detector for single-sideband reception. O. G. Villard and D. L. Thompson, June, 1948, p. 11.
A " Q5-er " for BC-348 owners. June, 1948, p. 59.
Selectable single-sideband reception simplified. J. L. A. McLaughlin. April, 1948, p. 11.
Selectivity in S.S.S.C. reception. O. G. Villard. April, 1948, p. 19.
An answer to N.F.M. reception. L. H. Allen. February, 1948, p. 28.
The " Q5-er." P. S. Rand. December, 1947, p. 18.
Building a code-practice receiver. R. M. Smith. December, 1947, p. 28.
Triode mixer *v.* pentode amplifier. J. Tannenbaum. November, 1947, p. 30.
Better N.F.M. reception with A.M. receivers. E. Harrington and W. Bartell. November, 1947, p. 38.
More on the BC-348-series receivers. November, 1947, p. 66.
Exit heterodyne QRM. J. L. A. McLaughlin. October, 1947, p. 13.
How sensitive is your receiver ? B. Goodman. September, 1947, p. 13.
Simple *L* and *C* calculations. J. Najork. September, 1947, p. 31.
The *Collins* 75A, September, 1947, p. 54.
Cathode-coupled converters for surplus receivers. J. H. Bender. August, 1947, p. 37.
The *National* NC-173. July, 1947, p. 47.
A new noise-reducing system for C.W. reception. D. L. Hings. June, 1947, p. 21.
The dialless converter. J. H. Creutz and D. F. McAvoy. June, 1947, p. 34.
Adapting the car radio to a converter. B. C. Barbee. June, 1947, p. 60.
The *Hallicrafters* SX-42. May, 1947, p. 48.
Modernising the old receiver. W. L. North. April, 1947, p. 54.
The old stand-by. G. D. Knipe. February, 1947, p. 42.
Converting the BC-348-Q. P. M. Kersten. January, 1947, p. 19.
Revamping the BC-342. September, 1946, p. 42.
An amateur-band eight-tube receiver. B. Goodman. August, 1946, p. 13.
Permeability-tuned oscillators. T. A. Hunter. August, 1946, p. 42.
Applying A.M.D. to the communications receiver. D. A. Griffin and L. C. Waller. August, 1946, p. 56.

Audio-modulated detection. D. A. Griffin and L. C. Waller. July, 1946, p. 13.
The *Hallicrafters* model S-40. July, 1946, p. 69.
The *Hammarlund* HQ-129-X. June, 1946, p. 24.
Noise limiting in C.W. reception. G. Grammer. May, 1946, p. 13.
New tuning system for the amateur receiver. W. J. Halligan and N. Foot. May, 1946, p. 18.
A band-pass 28-Mc. converter. B. Goodman. April, 1946, p. 44.
Panoramic reception, 1946. J. R. Popkin-Clurman and B. Schlessel. March, 1946, p. 22.
A 4·3-Mc. F.M./A.M. I.F. and audio amplifier. J. W. Brannin. March, 1946, p. 51.
Simple capacitance and inductance measurements. T. A. Gadwa. March, 1946, p. 71.
A 28-Mc. receiver/converter. B. Goodman. February, 1946, p. 17.
A 21-tube all-purpose receiver. J. Marshall. November, 1945, p. 31.
Graphical solution of bandspread problems. V. S. Buccicone. June, 1945, p. 42.

SHORT WAVE MAGAZINE

Signal-to-noise ratio and receiver gain. E. Dandy. October, 1949, p. 584.
Double superhet for ten. A. B. Wright. August, 1949, p. 418; September, 1949, p. 514; October, 1949, p. 594.
Noise limiter circuit. E. J. Williams. August, 1949, p. 440.
Broad-band converter for the BC-312. K. Bunston. June, 1949, p. 270.
Comparing receiver performance. W. J. Crawley. June, 1949, p. 286.
Radiovision " Commander." March, 1949, p. 55.
Effective audio filter. J. N. Walker. November, 1948, p. 630.
Modifying the HRO. I. E. Hill. October, 1948, p. 536.
BC-453 for better selectivity. F. E. Wingfield. September, 1948, p. 456.
Regeneration in superhets. A. D. Odell. May, 1948, p. 167.
Wide-band R.F. pre-amplifier. G. A. Hume. April, 1948, p. 88.
Improving the IF/AF amplifier. W. J. Crawley. January, 1948, p. 660.
More on the Synchrodyne. D. G. Tucker and J. F. Ridgway. December, 1947, p. 598.
Maximum signal-to-noise ratio. G. H. Cox. December, 1947, p. 602.
Local oscillator design. R. Knowles. November, 1947, p. 537.
The BC-348 modified. J. Hum. November, 1947, p. 540.
Amateur band receiver design. W. J. Crawley. August, 1947, p. 336.
The EF50 TRF receiver. J. Hum. July, 1947, p. 279.
Muting the AR88. N. P. Spooner. July, 1947, p. 282.
Another noise limiter design. J. H. Clarke. July, 1947, p. 293.
The *Radiovision* 5/10 expander. July, 1947, p. 304.
Noise limiter for the HRO. J. Ostle. June, 1947, p. 218.
The *Eddystone* 640 receiver. June, 1947, p. 232.
Modifying the Type 21 receiver. V. Stebbings. May, 1947, p. 142.
The Synchrodyne receiver. D. G. Tucker. May, 1947, p. 155.
A 5–10 converter for the R.1155. G. Elliott. April, 1947, p. 107.
A noise-silencer unit. W. R. Luckett. March, 1947, p. 21.
More on the EF50 receiver. J. Hum. December, 1946, p. 609.
Home-constructed communications superhet. A. B. Wright. October, 1946, p. 470; November, 1946, p. 546.
The *Eddystone* 504 receiver. September, 1946, p. 420.

High-performance TRF receiver. J. Hum. August, 1946, p. 340.
Principles of short wave reception. A. A. Mawse. March, 1946, p. 15; April, 1946, p. 78; May, 1946, p. 145.

WIRELESS WORLD

Eddystone Model 680. September, 1949, p. 335.
Is discriminator alignment so difficult? A. G. Crocker. September, 1948, p. 312.
High-level detection. W. MacLanachan. September, 1948, p. 330.
Series capacitor heater circuits. A. W. Stanley. September, 1948, p. 332.
The Synchrodyne. "Cathode Ray." August, 1948, p. 277.
Why align discriminators? T. Roddam. July, 1948, p. 240.
Single sideband selector. A. Dinsdale. July, 1948, p. 244.
Coil-pack modification. L. Miller. February, 1948, p. 59.
Quality superheterodyne. S. A. Knight. December, 1947, p. 472.
Tuned A.F. filters. H. E. Styles. July, 1947, p. 242.
Designing an F.M. receiver. T. Roddam. April, 1947, p. 143; June, 1947, p. 203.
A.V.C. calculations. S. W. Amos. February, 1947, p. 46.
Noise factor. L. A. Moxon. December, 1946, p. 391; January, 1947, p. 11; May, 1947, p. 171.
Noise limiters. H. B. Dent. December, 1946, p. 397.
Ex-R.A.F. communication receiver (Type R1155). "Ex-Signals." July, 1946, p. 212.
A.C./D.C. voltage dropping. July, 1946, p. 236.
Negative feedback. "Cathode Ray." February, 1946, p. 41; March, 1946, p. 76.
Ventilating problems. W. Tusting. March, 1946, p. 72.
Amateur communication receiver. H. B. Dent. February, 1946, p. 36.
Army set—Type R107. August, 1945, p. 235.
Modulated test oscillator. K. W. Mitchell. February, 1945, p. 56.
High selectivity at audio and intermediate frequencies. E. Lloyd Thomas. June, 1944, p. 175.
I.F. transformer design. J. McG. Sowerby. March, 1944, p. 82.
"Superhet–Straight" switching. P. A. Shears. January, 1944, p. 24.
Designing superhets. J. E. Haworth. June, 1943, p. 163.
Effect of a screening can on the inductance and resistance of a coil. J. McG. Sowerby. November, 1942, p. 254.
Variable selectivity IF amplifiers. J. E. Varrall. September, 1942, p. 202.

INDEX

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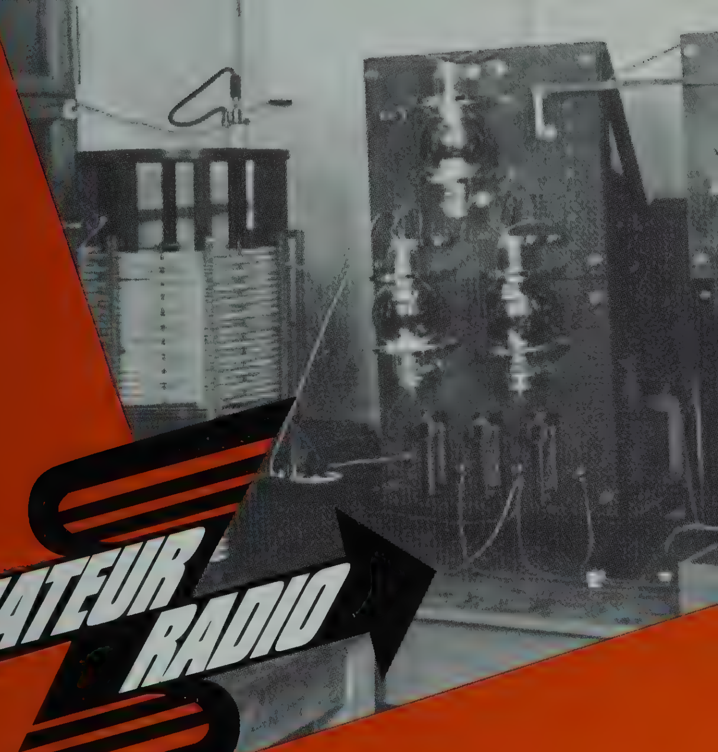

AMATEUR RADIO

SIMPLE TRANSMITTING EQUIPMENT

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CONTENTS

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FOREWORD

THE fact that it is possible to communicate over great distances with the aid of extremely simple equipment no doubt explains why so many men and women have joined the Amateur Radio movement during recent years. There is something fascinating in the knowledge that a small two-valve transmitter, costing only a few pounds, is capable of radiating a signal to the far corners of the earth.

Although contacts with stations located in distant lands provide most of the big thrills to be found in Amateur Radio, real enjoyment can be obtained from conversations with stations much nearer home.

In this book, therefore, the authors have described simple transmitters of different types each of which was designed to perform a specific type of work. For instance the " top band " transmitter is ideal for round-Britain contacts whilst the 25-watt set is capable under good conditions of radiating a useful signal into such distant lands as Australia and New Zealand.

It is appropriate that the authors should emphasise the necessity of employing an effective aerial system, for it is well-known that the most efficient transmitter will fail if it is operated with an aerial of poor design. The authors also emphasise the important part which frequency measuring equipment plays in the operation of short wave equipment; a simple frequency sub-standard similar to the one described will prove a boon wherever it is installed.

Each item of equipment referred to in the text has been designed to help the newcomer to Amateur Radio to achieve effective results for a modest outlay and, provided the appropriate specifications are followed as closely as circumstances permit, no difficulty should be experienced in reproducing results similar to those obtained with the original model. Incidentally the various components specified are all of British manufacture.

The publishers are indebted to Mr. J. N. Walker, G5JU, for furnishing full details of the " Two Valve Top Band " transmitter described in Chapter 3, and to Mr. F. Charman, B.E.M., G6CJ, for permission to reproduce certain material contributed by him to *The Amateur Radio Handbook*.

Readers are referred to companion publications in this series for full information on Short Wave Receivers, Valves, V.H.F. and Microwave Equipment. The booklet *Transmitter Interference* should also be consulted for details of suppression devices.

J. C.

CHAPTER 1 FUNDAMENTALS

THE first concern of a radio amateur should be to understand the principles of modern radio transmission. The mere generation and radiation of R.F. energy is not enough. His success or failure will be judged by the quality of his signals and by his ability to direct them towards those parts of the world with which he wishes to establish communication.

Illustrations in the radio journals tend to mislead the newcomer into supposing that the average amateur station consists of a mass of high-power equipment. In actual fact, most British stations occupy little more space than a table top in the living room, an attic or a cupboard under the stairs. Simple transmitters can, and do, give excellent results. The essential requirements are enthusiasm and patience in acquiring technical skill and knowledge. The broad outlines of transmitter and aerial design must, therefore, be discussed before we can pass to the details of the construction of simple but effective equipment for the amateur bands.

Let us first define what is meant by the term " good quality " when applied to a C.W. (telegraphy) signal. Firstly, the frequency of the transmission must be stable—primarily a function of the oscillator producing the R.F. carrier. Secondly, the keying must be clear-cut, free from chirp, clicks, and other extraneous noises. Thirdly, the note must not be modulated in any way by a poorly smoothed H.T. supply.

For the production of a C.W. carrier meeting the requirements set out above there are two basic oscillator arrangements which may be employed, namely, the quartz crystal oscillator and the master oscillator (M.O.) or variable frequency oscillator (V.F.O.).

When considering telephony transmissions, the necessity for a clean, stable carrier is every bit as important as with telegraphy. In addition, the modulation should be crisp, have no effect upon the stability of the carrier, and take up the least amount of precious ether space in crowded bands consistent with the transmission of an adequate range of frequencies for speech communication. Amateurs are concerned with communication; not with the transmission of music. High-fidelity modulation, with its requirement of a wide frequency range, is not only unnecessary, but is also productive of interference with stations operating on closely adjacent channels, and is out of place on the communication bands below 144 Mc/s.

To achieve long distance communication with low power, modulation should be fairly full, but, even on peaks it must not exceed 100 per cent., or the result will be severe interference both with other stations working on the band and with nearby broadcast receivers.

The Crystal Oscillator

This circuit, whereby a valve is made to oscillate at a frequency determined by the characteristics of a specially prepared quartz crystal, has both advantages and disadvantages for the amateur. To consider the advantages first:

(1) When properly adjusted the frequency remains stable to within a few cycles.

(2) It is a comparatively simple matter to obtain a good note, even when the oscillator is keyed.

(3) No frequency sub-standard is required by licence conditions beyond the certificate issued by the manufacturers of the crystal. Only simple frequency measuring equipment is, therefore, necessary.

The disadvantages are:

(1) The operating frequencies in any one band are limited to the number of crystals available: hence advantage cannot be taken of a clear spot in a crowded band.

(2) The cost of crystals will normally limit the newcomer to a few frequencies.

A suitable design for a crystal oscillator, based on one of the many equally effective circuits, will be found in a later chapter.

The Variable Frequency Oscillator

The current interest in the V.F.O. might tempt the newcomer to imagine that it is some new device, but in fact the basic principles of self-excited "master oscillators" are as old as the use of valves in radio transmitters. Due to the simplicity and clean note of the crystal oscillator, the V.F.O. fell into disfavour among radio amateurs and did not regain its popularity until just before the war, when congestion on the amateur bands emphasised the advantages of being able to change frequency rapidly. The development of new oscillator circuits also made it possible to achieve a degree of stability comparable with that of a crystal oscillator.

Some of the advantages of the V.F.O. compared with the C.O. have already been noted. But against these must be set:

(1) The low output of the V.F.O. if stability is to be maintained.

(2) The necessity of isolating the oscillator from the later stages in the transmitter, requiring the use of one or more buffer stages.

(3) The need to possess a frequency sub-standard with which the frequency of the transmitter can be measured to an accuracy of at least 0·1 per cent. in order to comply with the licence regulations.

(4) The power supply for a V.F.O. and its associated buffer amplifiers must possess extremely good smoothing and voltage regulation, particularly where the oscillator stage is keyed.

To sum up, the V.F.O. is considerably more complex, both in its construction and operation than the C.O., but, nevertheless, provided due care is taken in regard to the mechanical as well as to the electrical design, very satisfactory results may be obtained.

There are many popular types of V.F.O. circuits, but the one described on page 40 is well representative of modern practice. With careful construction, it will give excellent results without demanding an unreasonable outlay on components or advanced technical skill.

Between Oscillator and Aerial

There are several reasons why an oscillator, whether crystal or V.F.O., should not be employed to feed the aerial directly. Firstly, as has been mentioned before, good frequency stability and high output do not go together; secondly the aerial, being subject to wind and weather, does not present the unvarying load which is desirable for good oscillator performance.

Although some amateurs achieve excellent results with low-power single-stage apparatus, such equipment calls for considerable skill and experience. Newcomers to the amateur bands are advised to rely upon apparatus of more certain and predictable performance, at least in the initial stages.

The Power Amplifier

The valve stage supplying R.F. power to the aerial is termed the power

amplifier (P.A.), and may be driven directly by the oscillator if output is required on the same frequency. If a multiple or "harmonic" of the oscillator frequency is required, further stages, known as frequency doublers, may be employed between the oscillator and the P.A.

The possible efficiency of a P.A. stage, that is to say the ratio of the R.F. power output to the H.T. power input is, for a number of reasons, normally much greater than for an oscillator. Since the terms of the British amateur licence limit the H.T. power which may be used on the valve feeding the aerial, more R.F. power can be obtained if this valve is a P.A.

Valves employed as R.F. power amplifiers are operated under different conditions to those in audio frequency amplifiers. In the latter service the grid is biased negatively to the mid point of its anode voltage anode current curve, and the value of the signal voltage is never allowed to become large enough in the positive direction to cause the flow of grid current or large enough in the negative direction to reduce the anode current to zero. Either of these states would introduce severe distortion in the output unless a push-pull arrangement was used.

This mode of working is termed "Class A operation," and under these conditions the power supplied to the valve anode must not exceed the maker's rating for anode dissipation.

When a valve is employed as an R.F. amplifier its negative grid bias is increased to several times the voltage required to stop anode current flowing, even when the maximum anode voltage is applied. This is known as "working the valve beyond cut-off"—"cut-off" being that grid voltage which just stops the flow of anode current for any given H.T. voltage

An R.F. "driving" voltage is now applied to the grid, either directly from the oscillator or *via* an intermediate stage. The amplitude of this input voltage is such that its positive peaks not only overcome the negative bias on the grid of the P.A. but provide a positive grid voltage for part of the time, so that a high anode current will flow during these periods. This is known as "Class C operation."

It must be remembered that each positive peak of the grid drive is followed by an equal and opposite negative peak. However, this negative voltage has no effect on the P.A. valve because the grid is already so biased that the anode current is cut off, thus the anode current in the P.A. consists of successive short pulses of high current interspersed with longer periods of zero current. It is these rest periods which permit the valve to be operated in a more violent manner than would be possible if the valve was passing current continuously.

If a tuned circuit resonant to the frequency of the R.F. input is connected in the anode circuit of the P.A., then as each successive pulse of current flows through the valve this tuned circuit will be forced into oscillation. A tuned circuit possesses what in mechanics would be termed inertia, and oscillation continues for a period after the pulse has ceased. The action is, therefore, somewhat akin to the pulses transmitted to a clock pendulum; each successive pulse of anode current helping to maintain steady oscillations in the resonant circuit.

The aerial is coupled to this anode circuit, and draws off and radiates a proportion of the R.F. power. The maximum permissible power which may be supplied to the valve is, therefore, greater for Class C than for Class A by the amount of this R.F. radiated power.

The Frequency Doubler or Multiplier

The circuit of a frequency multiplier is similar to that of a P.A., but the operating conditions are a little different. Even more negative bias is applied to the grid, and the R.F. driving voltage is correspondingly increased, with the result that the pulses of anode current become of greater intensity, but last for a shorter period. The tuned circuit in the anode is made to resonate at a multiple of the frequency of the drive, the sharp pulses maintaining oscillation within the tuned circuit at the higher frequency.

The efficiency of such a stage is lower than that of a P.A., and decreases as the order of the harmonic rises. Generally speaking a multiplication of 4 or 5 times is the maximum which can be obtained with one stage.

In later chapters will be found designs for three transmitters employing crystal oscillators. In two of the designs, provision is made for operation on twice the crystal frequency by using the P.A. as a frequency doubler. This is a compromise between efficiency and simplicity, but is quite satisfactory in practice. In the 25-watt transmitter, frequency multiplying stages are included, so that output may be obtained at full efficiency on any one of five bands with only one crystal.

CHAPTER 2 AERIALS

Introduction

THE design of the aerial system is of the utmost importance in an amateur transmitting station, for if it has not the ability to radiate usefully, the power supplied by the transmitter is wasted. The difference between a good and a bad aerial system can be as great as a hundredfold difference in transmitter power. The efficiency of a transmitter or receiver can be observed quite easily, but the radiation from an aerial cannot be seen, and its efficiency can often be determined only by patient trial over a long period, based on reports received from other stations. An understanding of the basic principles of aerial design will, therefore, save the newcomer much time and trouble.

Briefly the problem is to transfer the R.F. power output of the transmitter into the aerial system, whence it must all be radiated in the form of ether waves, without local absorption of the radiated energy. Sometimes uniform radiation in all directions is desired, often it is intended to be directional; an aerial may be required to send signals over short or long distances, or both; the same aerial may have to work on one or on several amateur bands. In practice, an amateur's choice of aerial is often largely determined by the space available.

In the consideration of aerials it is more convenient to think in terms of "wavelength" rather than "frequency." It is usual to talk of the "frequency" used in a transmission, but of a "half-wave," or "full-wave" aerial. The relationship of frequency to wavelength is explained later under the heading "Resonant Length."

Marconi and Hertz Aerials

Aerial systems have tuning properties, even as the tuning circuits of the transmitter and receiver, and are, in effect, open oscillating circuits, the only

undamental difference being one of scale. In order to radiate well the aerial must be of appreciable dimensions compared with the wavelength of operation. The tuned circuits of the transmitter itself are small and confined, and, therefore, do not radiate appreciably.

Simple aerials fall roughly into two classes, the *Marconi* and the *Hertz*. The difference between them is a simple one, but it is of fundamental importance in understanding the design and operation of transmitting aerials.

In the Marconi aerial a wire of indefinite length is used in conjunction either with an earth connection or with a counterpoise. A counterpoise is a wire or wires erected below, but not necessarily immediately underneath, the aerial, about 5 or 6 feet from the ground and well insulated from it. In appearance, the counterpoise resembles a low aerial. It is employed in place of the more usual earth connection, and forms part of the resonant aerial system.

The Marconi aerial, whether employing an earth or a counterpoise, is brought into resonance with the transmitter frequency by means of a coil connected in series with these two elements. This coil provides the means of coupling to the transmitter, and is tuned with a condenser either in series or parallel, depending upon the wavelength used and the dimensions of the aerial system.

At one time the radio enthusiast took immense care over his earth connection. With the crystal receiver and the early valve sets an efficient, low resistance earth was of prime importance in securing the best performance. With modern receivers, losses resulting from a poor or non-existent earth contact pass almost unnoticed, and the making of a good earth has become a lost art.

It is essential that the earth connection employed in conjunction with a Marconi aerial should be of low resistance. The earth rod, either copper or brass, should be of substantial proportions—not less than, say, ¾ to 1 in. in diameter—and some 4 ft. long, driven into ground which will retain as much moisture as possible. In situations where the soil is abnormally dry, or where it is impossible to use a rod of the length indicated, two or three shorter rods should be driven into the ground a foot or two apart. The earth lead should be of substantial gauge wire as short and direct as possible, preferably with an insulated covering. The lead should always be soldered to the rod or rods. Where there are any doubts regarding the quality of the earth connection a counterpoise should be used instead.

The operation of the Hertz aerial is based on the fact that the natural wavelength to which any wire will tune depends directly upon its length. If a suitable length is chosen the radiator is thus self-tuned and no earth or counterpoise is necessary. A Hertz aerial possesses the advantage that it can be placed where it is less disturbed by the absorption effects of the earth, buildings, etc., and so is normally more efficient than the Marconi type.

Short-wave aerials are usually of the Hertz type, because of their greater efficiency, whilst the lengths of wire involved are convenient. Above about 100 metres the limitations of space usually force the amateur to use a Marconi aerial which can be made shorter than the corresponding Hertz aerial.

Resonant Length

The frequency of resonance of an aerial wire is related to its length, and is approximately half the length of the wave concerned, or any multiple of a half-wavelength. It is often more convenient, therefore, to think in terms of wavelength rather than in frequency, which is the reciprocal of wavelength, when dealing with aerial systems.

The velocity of a radio wave in space is 300 million metres (186,000 miles) per second. The number of complete oscillations in this time, or distance, is the frequency of the wave; thus the distance divided by the frequency gives the wavelength. (For example, 14 Mc/s. equals a wavelength of 21·4 metres.) When travelling on a wire, however, the velocity of the wave is slightly reduced if there is any radiation taking place, and so a half-wave aerial for 14 Mc/s. is not quite half of 21·4 metres, but a small percentage less than this figure. The correction in practice is usually about 5 per cent. In a wire on which several half-waves exist this correction is only applied to the end quarter-waves, so (including a factor for conversion from metres to feet) the formula for an aerial of resonant length for any frequency becomes:

$$\text{Length} = \frac{492\,(n - \cdot 05)}{f}\ \text{feet.}$$

where n is the number of complete half-waves on the aerial, and f is the frequency in megacycles per second. An accuracy of the order of 1 per cent. is usually satisfactory when calculating resonant lengths.

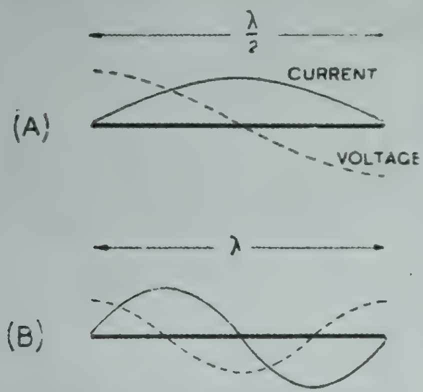


Fig. 1.
Standing waves in resonant aerials, showing voltage (broken line) and current (full line) variation along the wire. (A) Half-wave or fundamental frequency. (B) Full wave, or 2nd harmonic. The symbol λ is generally used to denote a wavelength.

In Fig. 1 the wires are shown in resonance with the applied oscillations. If they were not cut to correct lengths the standing waves could not develop fully and radiation would be reduced; it would then be necessary to alter the length of the aerial to correct this condition. If it were not possible to do this directly, then it would be necessary to *load* the aerial with capacity or inductance. Examples will be seen later.

Radiation Resistance

It is not within the scope of this book to attempt to describe how radiation of radio energy from an aerial takes place, but just as the power consumed in any electrical circuit is a function of the current flowing through the resistance of the load, so an aerial is said to possess a *radiation resistance* in which the current flowing in the aerial is dissipated. The radiation resistance may be determined at any point in the aerial, but it is usual to refer to a point of maximum current which, in the case of a half-wave aerial, is at the centre, and has a value of approximately 75 ohms.

From the formula: Watts $= I^2R$, a current of 1 A. in the centre of such an aerial would, therefore, represent about 75 watts of radiated power.

The Marconi Aerial

On the 1·8 Mc/s. (160 metre) band and often on the 3·5 Mc/s. (80 metre) band it is seldom practicable to erect a Hertz aerial within the confines of the normal garden. Particularly is this true on 1·8 Mc/s. where a half-wave aerial would be approximately 265 ft. long.

It is on these lower amateur frequencies that it becomes necessary to employ the Marconi aerial in conjunction with an earth or a counterpoise. As previously mentioned, the Marconi may be of an indefinite length, although the shortest stretch of wire which will permit maximum current at the base together with maximum voltage at the top—thereby ensuring reasonable efficiency—is one quarter-wavelength.

The question now arises as to how the R.F. power from the transmitter may best be transferred to the aerial. In the case of a Marconi approximately one quarter-wavelength long this is accomplished by means of a coupling coil connected between the lower end of the aerial wire and the earth or counterpoise. This coil will, in effect, form part of the aerial, thus tending to increase its effective length. In order that the centre of the coupling coil shall be the point of maximum current in the system, means must be provided

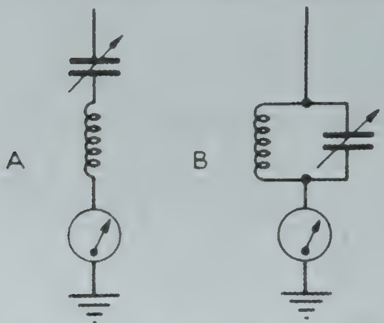


Fig. 2.
Marconi Aerials.—(A) Shows feed method when aerial is about quarter-wave long. (B) Shows feed method when aerial is less than quarter-wave long, or when it is an odd multiple of quarter-waves.

to cancel out the loading effect of the coil. This can be accomplished by placing a variable condenser in series with the coil, as shown in Fig. 2 (A). The condenser also provides a convenient means of tuning the aerial system to resonance. A useful guide to the size of coil required is that it should contain about half the quantity of wire which would be needed to make the aerial system up to a complete half-wavelength. For this purpose an earth should be considered as a quarter-wavelength. The condenser should have a maximum capacity of 250 to 300 $\mu\mu$F. to allow of adequate adjustment. With an aerial of less than a quarter-wavelength the method of tuning should be as shown in Fig. 2 (B), where it will be seen that the coupling coil and condenser are in parallel to increase their loading effect. A similar value of condenser to that employed in the series connection would again be suitable, while the size of the coil, depending as it does to some extent upon the actual length of the aerial, should be rather larger than with series tuning, and have tappings brought out every few turns so that its effective inductance may be adjusted as required. In the case of an aerial which is an odd multiple of a quarter-wave in length the circuit in Fig. 2 (B) should be employed.

It will be noticed that in both these diagrams an R.F. ammeter is shown in series between the coupling coil and the earth or counterpoise. A 0–1 A.

meter—either hot-wire or thermo-couple type—will serve as a convenient indicator of resonance. The reading obtained will depend upon the position of the meter in relation to the point of maximum current of the whole aerial system. An increase in current with other factors unchanged will indicate an increase in aerial power. This point is mentioned because of the frequent discussions which take place among amateurs regarding the aerial currents they obtain, without any reference being made to the actual length of their aerial *systems*—not the length of the aerial alone—and the position of the R.F. ammeter in relation to the voltage and current distribution.

Where, due to lack of space, it is impossible to get the whole length of aerial wire required between the available supports, it is permissible for 20 or so feet of the far end of the aerial to be run at an angle to the main wire, or even to hang vertically downwards. Since this end of the aerial is a point of high R.F. voltage, adequate insulation must be provided. Steps should also be taken to prevent this portion of the aerial from blowing about in the wind and touching nearby objects.

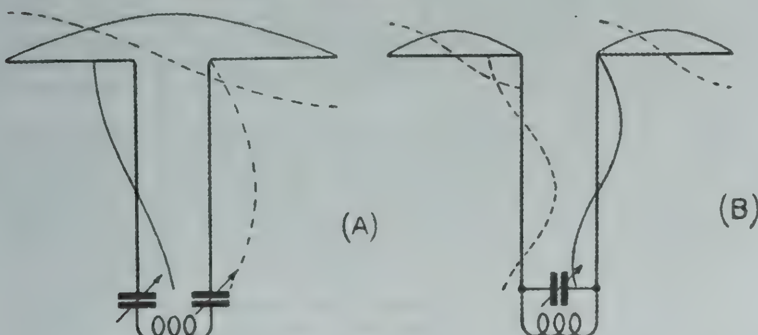


Fig. 3.

Centre-fed Aerials using tuned feeders. The standing waves are shown, voltage distribution by the broken line, and current by the full line. (A) is a current fed half-wave aerial with series tuning at the transmitter, and (B) is a voltage fed full-wave aerial with parallel tuning. The choice between series and parallel is governed by the length of the feeder. See the chart of Fig. 5.

The Hertz Aerial

It has already been stated that the shortest length of wire which will resonate without loading to a given wavelength is the half-wave; this is the simplest form of Hertz aerial. Lengths of wire which represent any multiple of half a wavelength, at the frequency concerned, are also self-resonant, and it will be apparent, therefore, that an aerial which is in half-wave resonance at a certain frequency will exhibit full-wave resonance at twice that frequency, and carry two full waves when the original frequency is increased fourfold. It is this property which enables one length of wire to be employed as an efficient radiator on several harmonically related amateur bands.

Feeding the Aerial: Feeders or Transmission Lines

By the use of feeders or transmission lines, the energy of a transmitter may be carried a considerable distance without appreciable loss providing there is no radiation or leakage of power from these feeders. Such devices

make it possible to place the radiating element in a position where it may be used to maximum advantage. For example, a Hertz radiator is only 33 ft. long on 14 Mc s., yet it could be placed as high as 100 ft. and supplied efficiently with R.F. power. If on the other hand, no feeders were used, it would be necessary to bring part of the radiator to the transmitter, and so most of this valuable height would be lost. Part of the radiated power would also be lost in the building, and it would be difficult to arrange for signals to be sent in any desired direction. The use of a feeder thus allows us to control more easily the behaviour of the aerial which may be placed and oriented as desired.

There are two main types of feeder. The first consists of a pair of parallel wires a few inches, or fractions of an inch, apart. Such an arrangement is often called a *twin* or *balanced feeder*. At any point along the feeder the currents in the two wires are equal in value but opposite in sign, and as the wires are close together the radiation from them cancels out. This type of feeder may be operated in two ways. If equal and opposite standing waves are distributed along its length it is called a tuned feeder, and the current and voltage is as shown in Fig. 3. This arrangement requires to be tuned at the transmitter end so as to accommodate a definite number of quarter-waves. A device of this type is known to amateurs as a "Zepp feeder," and an aerial so fed as a "Zepp aerial," a name which dates from its original use in Zeppelin airships.

It is possible, on the other hand, to arrange the feeder so that current and voltage are distributed uniformly all along it, with no standing waves, and in this case the feeder is said to be *matched*. The matching is done by making the load at the far end assume such a value that it absorbs energy just as fast as the feeder supplies it, so that there can be no reflection, as in the case of the aerial which is open at the end. It is possible to utilise transformer action to match the radiation resistance of the aerial to that of the feeder, or in some special cases *vice versa*.

A matched feeder is always preferable to a tuned one from the point of view of efficiency. In commercial practice matched feeders are commonly made 1,000 ft. long without serious loss, but an unmatched line should not be more than about one wavelength long. In certain circumstances, however, the Zepp or tuned feeder has many advantages, particularly when it is desired to operate the same aerial on several bands.

The second type of feeder also has two conductors, but one conductor now forms a tube round the other and is spaced from it with suitable insulating material. Flexible feeders of this type, known as concentric or co-axial feeders are available for amateur use, but as they operate with the outer conductor connected to earth they are unbalanced, and, therefore, unsuitable for direct connection to the centre of a half-wave *Hertz* aerial—or *dipole* as it is sometimes called—without the addition of a balancing device.

The Zepp

The general appearance of this popular aerial is shown in Fig. 4 (A). The horizontal span can be half a wavelength or any multiple of half a wavelength long at the frequency to be transmitted. One of the twin feeders (L) is connected to the aerial and the other left unconnected, but well insulated. These wires (14 or 16 S.W.G. copper) are usually kept about 6 in. apart by the use of spreaders made from waxed wood, glass, polythene or ceramic, spaced at intervals of 3 ft. or so along the feeder. Suitable spreaders of this type are readily available, whilst lengths of feeder, with spreaders already fixed in place, may be purchased complete.

At the transmitter end the feeders may be either series tuned (A) or parallel tuned (B); the choice is determined for any particular frequency band by the length (L) of the feeder. If there is a current maximum at the transmitter end, *i.e.* when the feeder is an *odd* number of quarter-waves long, then series tuning is needed, and the current in the R.F. meters may be as high as 1 A. for 25 watts input. With a voltage maximum (even number of quarter-waves on the feeders) the current is very low, and parallel tuning is necessary. The latter condition should be avoided if possible, as it is then more difficult to adjust the coupling. This may be done by increasing the feeder length a few feet. The extra length may, if necessary, be accommodated by making the feeder run from aerial to transmitter less directly, but in so doing sharp bends should be avoided. The chart in Fig. 5 shows how to locate the standing waves for various frequency bands and lengths of feeder, and indicates the type of coupling circuit required.

The best adjustment for a Zepp aerial does not give—as is often supposed—exactly equal currents in the two feed wires, but can be determined as follows. First disconnect the radiator, and hoist the feeder into position. Connect the coupling circuit and couple it very loosely to the transmitter. Now tune to resonance. If series tuning is used, work the two condensers together for equal currents at resonance. The P.A. should then draw maximum anode current and the resonance should be sharp. Note the condenser readings carefully. The radiator should be cut a little longer than the correct length for the required harmonic. Connect it in position, hoist the aerial and try the tuning of the coupling circuit, increasing coupling as necessary. If the tuning does not agree with the previous figure, shorten the top an inch or two at a time, until the original condenser readings are obtained. Then, without altering the tuning, increase coupling until the meters show a maximum current in the feeder, or, in the case of the parallel connection being in use, until a neon lamp indicates maximum voltage across the coupling coil.

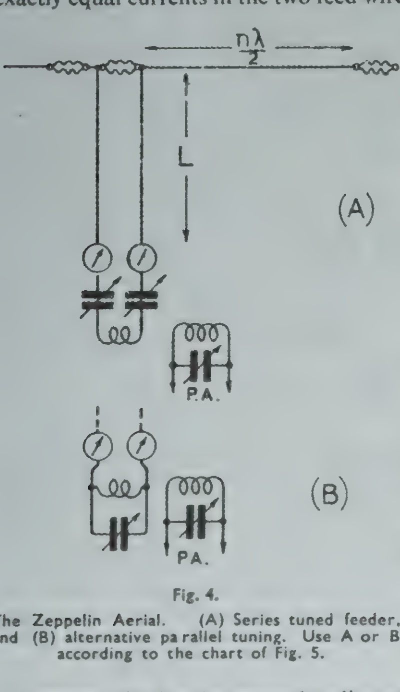


Fig. 4.

The Zeppelin Aerial. (A) Series tuned feeder, and (B) alternative parallel tuning. Use A or B according to the chart of Fig. 5.

The system will then be correctly adjusted although the feeders will probably not draw exactly equal currents. As previously stated, the top may be any number of half-waves long, but the optimum adjustment just described may only be made for one band. If other bands are to be used the adjustment should be made for the band on which highest efficiency is required; on other bands tune for equal currents in the two feeder wires.

The most suitable length of feeder for use with the Zepp aerial is the

quarter-wave, permitting the employment of series tuning, but this length becomes a half-wave on the second harmonic (next band higher in frequency). In multi-band operation, therefore, it will often be necessary to change from series to parallel tuning when changing bands.

For the lower frequency bands, where the top is less than half a wavelength long, Zepp operation is not possible, but the free feeder may be left unconnected at the transmitter end and the remaining feeder wire and the aerial or " top " used as a Marconi or an " end-fed " aerial.

The effective length of any aerial should be considered as extending to the *centre* of the loop of wire which passes round the insulator.

The End-fed Aerial

This is the simplest form of Hertz aerial inasmuch as it requires no feeders, but for that reason the permissible height, particularly on the higher frequency

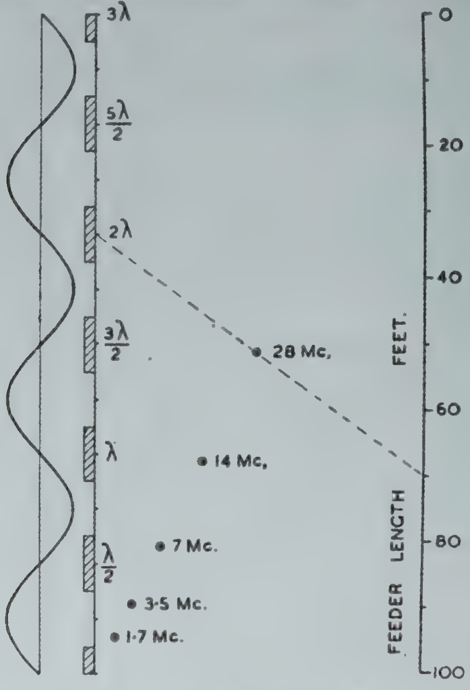


Fig. 5.

Standing Wave Chart for Tuned Feeders. The shaded scale is marked in quarter wave units. A line from the feeder length point through the frequency point cuts the scale at the required number of quarter waves, e.g. 70 ft. feeder is 8 quarter waves on 28 Mc/s.

Voltage Fed—Centre-fed top an even number of half-waves long, or Zepp any length—curve shows current distribution on feeder. Shaded regions need parallel tuning, and unshaded, series tuning (Fig. 3).

Current Fed—Centre-fed top an odd number of half-waves long—curve shows voltage distribution. Shaded regions need series tuning, and unshaded, parallel tuning.

bands, is limited since one end of the radiator has to be brought to the transmitter for coupling purposes. Suitable coupling arrangements are shown in Fig. 6, where at (A) the aerial coil is inductively coupled to the P.A. anode circuit, and at (B) coupled to the same circuit through the intermediary of a "link line," which will be more fully described later. Direct inductive coupling, while electrically efficient, is somewhat difficult to accomplish mechanically, as the position of the aerial coil has to be adjustable in relation to the fixed P.A. coil. With "link coupling" the aerial and P.A. coils may be separated by any convenient distance, *i.e.* the aerial coil could be situated near to where the aerial enters the transmitter room, while the P.A. coil would be in the transmitter, perhaps on the opposite side of the room. This gap can be spanned by a length of low impedance twin or co-axial feeder, to each end of which is connected a few turns of wire, coupled to the respective tuned circuit. The coupling between aerial and P.A. coils may be varied by

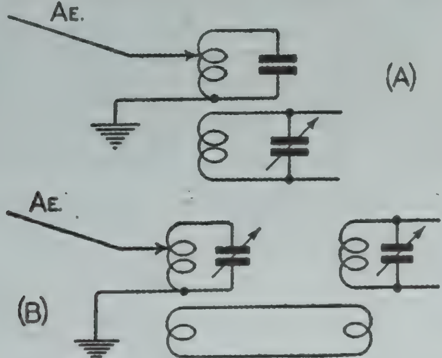


Fig. 6.

Coupling an End-Fed Aerial. (A) Loose coupling, and (B) link coupling.

adjusting the position of *either* of the link coils relative to the tuned coil to which it is coupled. Each link coil should be of the same diameter as the coil to which it is coupled and should consist of one to one-and-a-half turns of wire on the 14 and 28 Mc/s. bands, two to three turns on the 7 Mc/s. band, and three to four turns on 3·5 Mc/s. The final number of turns for optimum performance depends upon several factors, and is best found by experiment, but the number at either end of the link line should always be the same. In all cases the link coil *must* be coupled to the *earthy* end of the P.A. and aerial coils. With the P.A. coil this would normally be the end of the coil to which the H.T. is fed, or, in the case of a coil which is centre tapped, or where a push-pull P.A. is in use, to the centre of the coil.

When a length of co-axial feeder is employed for the link-line the outer casing should be earthed by connecting it at one end to the chassis of the transmitter, and at the other directly to the earth lead from the aerial coupling coil. With a two wire link-line the centre point of the link coil should be taken to earth in order to guard against the transference of harmonic energy to the aerial *via* the stray capacity between the link coils and their associated tuned circuits.

Under no circumstances should the aerial be tapped either directly or through a condenser on to the tank coil of the P.A. The former method places the full voltage of the H.T. supply on the aerial wire; not only is this practice extremely dangerous, but it usually contravenes the terms of the amateur licence. Even with the latter system there is the possibility of causing severe broadcast and television interference as well as radiating a large amount of energy on the harmonic frequencies with a consequent lowering of efficiency.

Connecting an aerial to a tuned circuit should not result in any appreciable detuning of the latter, because a resonant aerial, or aerial and feeder system, presents a purely resistive load, the only effect of which is to flatten the tuning of a circuit to which it is coupled. This fact provides a valuable indication of whether an aerial system is in resonance. For if the aerial is too short, *less* capacity will be required to maintain resonance, and if too long *more* capacity will be needed when compared with the unloaded circuit.

The Dipole

Next to the end-fed aerial, the dipole is probably the simplest aerial to erect but it suffers from the disadvantage that it is only effective on the frequency band for which it is cut. Where, however, activity is to be confined to one band, or where space permits the erection of several aerials for different bands, the dipole provides a most satisfactory radiating system.

The general appearance of the centre-fed dipole aerial is shown in Fig. 7, and for the purposes of this description only low-impedance balanced feeders

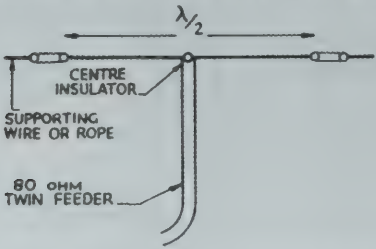


Fig. 7.
Dipole aerial with low impedance balanced feeder.

will be considered for connecting it to the transmitter or receiver. Other methods of feeding may be employed, but on the score of efficiency, cost, and ease both of handling and of coupling to the transmitter, the low impedance twin feeder offers many advantages.

The length of the horizontal span will depend upon the frequency to be employed, and the approximate lengths of wire required for the various amateur bands are set out at the end of this chapter. It should be noted that these lengths are those of the wire itself, and do not include the space at the centre of the aerial where the feeder is connected. In other words the overall length of an aerial 33 ft. long will be 33 ft. *plus* the spacing at the centre, or about 33 ft. 1½ in. Special insulators may be obtained for use at this point, but the suggested figure of 1½ in. is not unduly critical.

Suitable feeder cable, with an impedance of approximately 80 ohms, is available from several manufacturers, and consists of a pair of closely spaced parallel copper wires moulded in a high-grade insulating material of oval section. Although mechanically strong, this cable is light and does not deteriorate with exposure to the weather over a lengthy period.

The lengths given in the table are calculated according to the formula on

page 9 but it must be stressed that these may not be exact for every location, since the effects of the height of the aerial above the ground and its proximity to buildings, trees, etc., may have to be taken into account. On the lower frequencies such factors do not have so great an effect as at higher frequencies. Where it is desired to have the aerial exactly resonant on any given frequency a certain amount of " cut and try " is inevitable. However, there is usually no call for extreme accuracy, particularly if it is desired to operate on different frequencies within the same band. An aerial cut for the centre of a band, or the centre of a particular portion of a band as, for instance the C.W. section of 28 Mc/s., will normally be found to perform satisfactorily over the entire band.

The Directional Effect of an Aerial

An aerial radiates and receives energy in certain directions better than it does in others, the effect depending upon a number of factors of which the

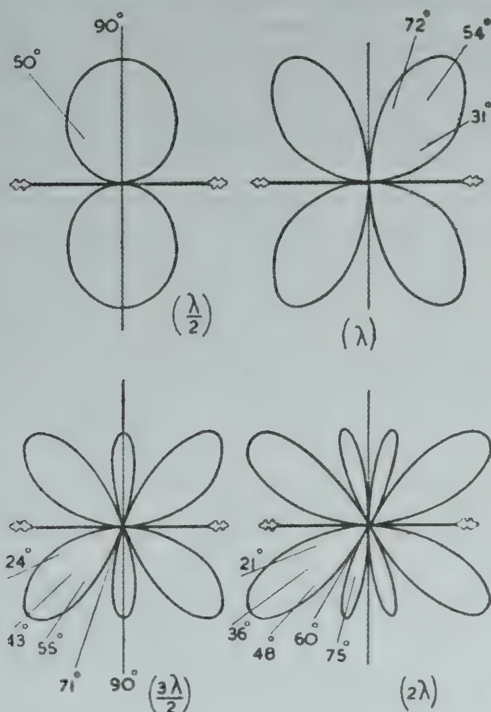


Fig. 8.

Polar Diagrams of Long Wires from λ/2 to 2λ in length. The angles of the main lobes and crevasses are shown, and also the angles for a 3 dB. loss in the main lobes. These diagrams are for free space: if the wire is horizontal the end crevasses tend to fill in giving end-fire radiation, mostly at high angles.

most important are the size of the aerial in relation to the wavelength in use, and its height above ground.

Unfortunately, it is seldom possible for the average amateur to place the aerial in just the position where it would send and receive signals as desired. Either the garden does not run in a favourable direction, or there may be obstructions, such as trees, buildings and overhead telephone and power wiring, to be considered.

Armed with some knowledge of how different types of aerial tend to radiate it is usually possible, however, to choose a suitable design to make the best use of the space and facilities available.

Polar Diagrams

In order to compare the directional performance of different aerials it is usual to employ a graphical representation of the radiation pattern known as a " polar diagram "; examples for various types of aerials will be found in Fig. 8. If an aerial is energised, and a suitable measuring device, such as a field strength meter, is so placed at various points around it that the reading on the instrument remains constant, then a line joining all these points would constitute the polar diagram of the aerial *in that particular plane*.

It is impossible, of course, to represent on the flat surface of a diagram something which is three dimensional, for it must be remembered that the radiation from an aerial wire occurs at all angles around the wire. If, for example, the wire is *vertical*, the polar diagram will be a circle with the aerial at its centre, but as the radiation is not equal from all parts of the aerial, the field-strength intensity would vary with different positions of the measuring instrument in a direction parallel with the wire.

Communication with stations outside the range of the ground wave is carried on by virtue of reflection of the waves from one or another of the layers of ionised gas which encircle the earth at certain heights, and to ensure that the wave travels to its destination in the smallest number of " hops "—thereby suffering the minimum of attenuation by absorption both in the layer and in the ground—the angle at which it first strikes the layer should be as small as possible. This means that for optimum long-distance results the radiation from the aerial should be confined within a very small vertical angle above the horizon. Such a result cannot be obtained with plain dipoles or with other simple aerials, but the addition of further elements for the purpose of modifying the vertical polar diagram is the basis of most directional arrays or " beams."

The design of such arrays is not a simple matter and they cannot be recommended for use by the newcomer until some experience has been obtained with simpler types. No attempt will be made, therefore, to describe their design or construction in this booklet.

It will be seen from Fig. 8 that whereas a half-wave dipole radiates best in a horizontal direction at right angles to the wire, aerials containing a number of half-wavelengths tend to radiate in other directions, the main lobes of energy coming more and more in line with the wire as its length is increased. This effect may be turned to advantage in locations where it is impossible to erect a half-wave aerial to give the required coverage.

If the height of an aerial above ground is small compared with the wavelength in use, interaction between the energy leaving the wire and that reflected from the surrounding ground will tend to increase the radiation at high angles at the expense of that at the more desirable low angles. It is better, therefore, in the interests of long distance communication, to erect the aerial at least one half-wavelength above the ground, in fact the higher

the better, provided that a height of three-quarters of a wavelength (½ plus ¼ wave) is avoided, as this exhibits many of the undesirable qualities of the lower aerial.

Aerial Erection

The wire should preferably be of hard-drawn copper not less than 16 S.W.G. Enamelled wire has the advantage of greater resistance to the effects of a corrosive atmosphere such as is found in large towns and near the sea. Stranded wire may be employed, but care must be taken to see that each strand makes good contact with the transmitter or receiver, and that there are no broken strands in the length.

Insulators having a long leakage path should be chosen, as there is then less chance of the aerial characteristics altering during wet weather.

An aerial must not be erected across a public highway without permission from the appropriate authorities, nor across power or telephone wires unless the owners are satisfied that their wires are fully protected from contact with the aerial should it or its supports collapse.

A substantial knife-switch should be arranged *outside the building* to connect the aerial to earth when it is not in use to ensure that static charges cannot build up on the wire.

Feeder spreaders for Zepp aerials may be made cheaply and efficiently from lengths of wooden dowel ⅜ in. or so in diameter provided they are well impregnated, and coated with paraffin wax. The procedure is as follows.

Cut the dowel into lengths equal to the required feeder spacing (normally 5 or 6 in.), drill a small hole close to the ends of each piece of dowel for the purpose of passing a short length of 22 S.W.G. wire, with which to bind the feeder wires across the ends of the spreaders. Soak the spreaders in hot—*not* boiling—paraffin wax until the wood has been well impregnated. This will be indicated when all bubbles cease. The wood is then removed, and the temperature of the wax allowed to fall slightly. The spreaders are again dipped into the wax and under these conditions a good coating will adhere and provide excellent protection against the weather for at least two years.

Rope for aerial support and for mast halyards may be rendered weather-proof, and prevented from contracting under damp conditions by the use of a similar method of waxing.

TABLE OF RESONANT AERIAL LENGTHS

Band. Mc/s.	½λ	1λ	1½λ	2λ	2½λ	3λ	3½λ	4λ
1·85	252·7	518·6	—	—	—	—	—	—
3·65	128·0	263·0	397·7	532·4	—	—	—	—
7·15	65·4	132·2	203·0	271·8	340·5	409·4	478·2	547·0
14·2	33·0	67·6	102·2	136·9	171·5	206·2	240·8	275·5
21·13	22·1	45·4	68·7	92·0	115·3	138·5	161·8	185·1
21·38	21·9	44·9	67·9	90·9	113·9	136·9	160·0	182·9
28·5	16·4	33·7	50·9	68·2	85·5	102·7	120·0	137·2
29·5	15·8	32·5	49·2	65·9	82·6	99·2	115·9	132·6

The frequencies shown in the left-hand column represent the centre points of each band, except in the case of the wider bands where two points are given. Aerial lengths are in feet.

CHAPTER 3 SIMPLE TRANSMITTERS

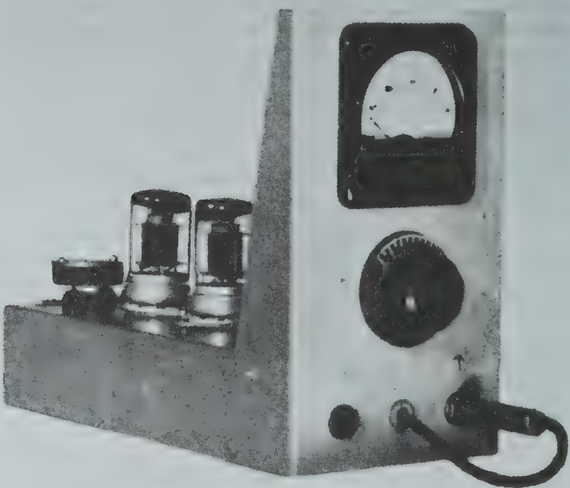
SIMPLE TWO VALVE TRANSMITTER FOR 1·8, 3·5 AND 7 Mc/s. OPERATION

THE transmitter to be described uses a minimum of components, is simple to construct and gives quite efficient results on the fundamental frequencies of the crystals used and only slightly less efficient results on the second harmonics.

Circuit Description

Using a 1·8 Mc/s. crystal, and the second valve (V2) as a Power Amplifier, inputs up to 10 watts can be obtained on this band. When a suitable coil is plugged into circuit, V2 can be operated as a frequency doubler and 10 watts obtained on the 3·5 Mc/s. band. Using a 3·5 Mc/s. crystal and operating V2 as a P.A. will give a second point in this band, while operating V2 as a F.D. will give a point in the 7 Mc/s. band. If a 7 Mc/s. crystal is used in the same way a second point in this band can be obtained, and also a point in the 14 Mc/s. band. Three coils only are necessary, as the 3·5 Mc/s. coil has been arranged to tune also to the 7 Mc/s. band.

The crystal circuit chosen (a Pierce oscillator) has the advantage of being extremely simple, since it requires no tuned circuit to make it oscillate. It has one disadvantage, however, in that the R.F. current flowing through the crystal is inclined to be rather high, and unless care is taken in



Photograph showing general view of the low power transmitter. The crystal and two valves can be seen on the chassis.

the design of the circuit a fracture of the crystal might result. In this transmitter the crystal current has been reduced to a minimum by careful choice of component values and the use of high efficiency valves. Crystals with 7 Mc/s. fundamental frequencies have proved quite successful, but when purchasing them it should be stated that they are for use in a Pierce oscillator circuit. *This point is most important.* Since the P.A. stage also employs a high efficiency valve which requires a very small amount of drive, this permits the use of very low power on the crystal stage, with a consequent further reduction of crystal current.

The anode circuit of the P.A. stage is parallel fed, that is, the H.T. supply is taken to the anode *via* a R.F. choke, and the coil is connected through a condenser to the anode. This allows one side of the tuning condenser to be connected to the chassis with a consequent simplificaton of construction, and removes H.T. voltage from the coil.

From an examination of the circuit diagram (Fig. 9) it will be seen that, in addition to a crystal socket, an alternative connection has been provided so that the output from a variable frequency oscillator may be used to drive the transmitter. It should be noted here, however, that unless there is a

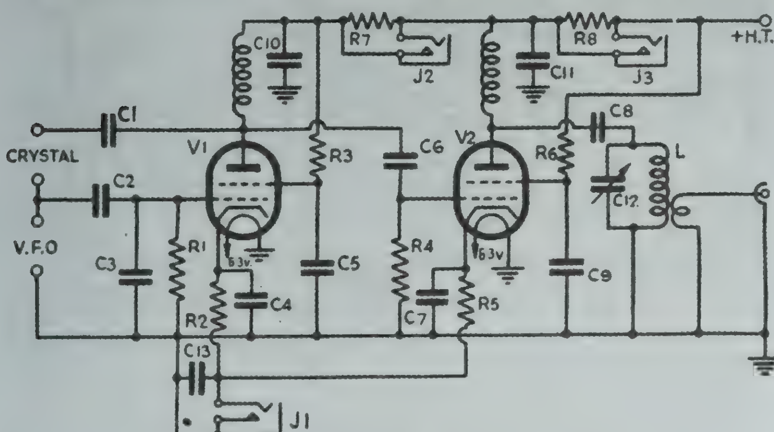


Fig. 9.

Circuit diagram of low power transmitter.

C1, 2, 8	·001 μF. *T.C.C. type CM20N.*
C3, 6	100 μμF. *T.C.C. type CC42h.*
C4, 5, 7, 9, 10, 11, 13	·01 μF. *T.C.C. type 543.*
C12	100 μμF. variable.
L	See coil table.
R1	50,000 ohms. *Erie type 8.*
R2	1,000 ohms. *Erie type 2.*
R3	250,000 ohms. *Erie type 8.*
R4	10,000 ohms. *Erie type 8.*
R5	300 ohms. *Erie type 2.*
R6	30,000 ohms. *Erie type 2.*
R7, 8	100 ohms. *Erie type 8.*
V1, 2	*Mullard QVO4/7.*
J1	Key Jack. *Igranic (Webb's Radio).*
J2, 3	Meter jacks. *Igranic (Webb's Radio).*

Other Components:

- Meter. *Pullin type 25 square flush.*
- Crystal sockets. *Q.C.C. type M.*
- Crystals. *Q.C.C.*
- Valve holders. *Belling Lee type L/500/C2/Ag.*
- Power input sockets. *Belling Lee type L1113.*
- Power input plug. *Belling Lee type L525.*
- Concentric output socket. *Belling Lee type L604S.*
- Coil socket. *Eddystone type 964.*
- Coil formers. *Eddystone type 538 (2) and 537 (1).*
- Tag strips. *Bulgin type T32 (4).*
- Chassis and panel. *E. J. Phillpotts Metalworks, Ltd.*
- R.F. chokes. *Eddystone type 737.*

considerable harmonic content in the output from the V.F.O. the transmitter will only function successfully on the same frequency band as that output.

Keying is achieved by opening the cathode circuits of both valves. If it is found that the key clicks are objectionable, one of the special key thump filters described in *Transmitter Interference* should be added.

The output is taken from the P.A. anode coil by means of a link coil wound on the same former, and thence to a concentric socket. From this a lead can be taken to any convenient point where an aerial coupling coil may be installed. A similar link coil—which should be movable—at the far end will complete the transfer of energy from the transmitter to the aerial. The movable link is necessary so that the correct coupling may be obtained to give optimum output.

A meter connected to a telephone plug may be inserted into either anode circuit for checking or tuning purposes. A low value resistor is connected

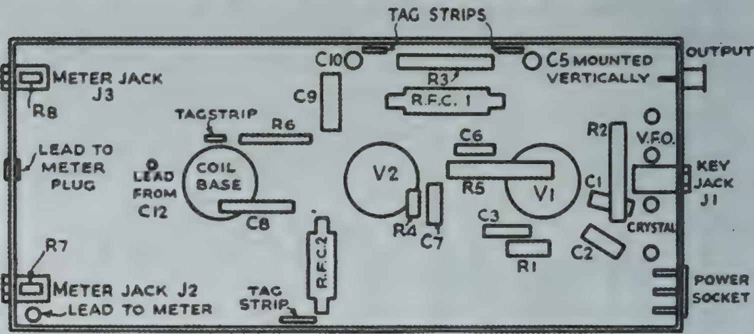


Fig. 10.
Sketch showing layout of components underneath the chassis. This drawing is not to scale.

across each jack so that as the plug is either inserted or withdrawn there is no break in the circuit.

Construction

The photograph (page 20) shows the general shape of the chassis and panel which are of 16 gauge aluminium. The chassis measures 10 in. × 4½ in. × 2 in. and the panel is 8 in. high, and shaped to give rigidity. On the panel is mounted the meter, and below that the tuning condenser (C12). The jacks J2 and J3 can be seen at the bottom with the meter lead and plug coming out in the centre.

The crystal and V.F.O. sockets are mounted at the rear of the chassis, then V1 and V2 and finally the coil socket. The rear edge of the chassis carries the power input socket, the concentric socket for the output, and the keying jack J1.

Fig. 10 shows the under-chassis layout of the components from which it can be seen that the larger resistances and the R.F. chokes are supported on small tag strips bolted to the chassis. The remainder of the smaller components are supported in the wiring.

Operation

Since there is only one tuning control, operation of the transmitter is very

simple, provided it is remembered that the 3·5 Mc/s. coil tunes to the 7 Mc/s. band with the variable condenser set near minimum, and to 3·5 Mc/s. when it is near maximum. In the same way the 7 Mc/s. coil tunes to 7 Mc/s. and 14 Mc/s.

Upon plugging in a crystal and coil for the band it is intended to use, and connecting a 250–300 volt power supply, the transmitter is ready for operation. A check on the anode and screen current of V1 (made by plugging the

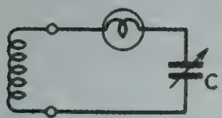


Fig. 11.
Circuit of simplified harmonic indicator. The coil may be interchangeable, and wound similarly to the transmitter coils.

meter into J2) should show a current of about 10 mA. The meter should then be plugged into J3, and the tuning condenser rotated to the region indicated in the coil table, when there should be a very pronounced dip. This is the point of resonance; it should, however, be emphasised that the transmitter must not be run for long periods tuned in this manner—*e.g.* without a load on V2. This is because, being a pentode, V2 will pass a high screen current when the anode current is at a minimum. Under load, of course, the valve operates with normal currents.

A 6 volt car bulb connected to the output link would form a suitable temporary load although this would be more effective if a tuned circuit were inserted first—say the aerial coupling circuit—the lamp then replacing the aerial as an absorber of power.

When an aerial has been connected to the coupling unit in a manner similar to that described in an earlier chapter, and the link coupling adjusted to give optimum output, the transmitter should be ready to go on the air.

Accessories

Two useful and simple pieces of gear will assist in tuning any transmitter, more especially those having a number of stages. They are the single turn loop and bulb, and the simple absorption indicator illustrated in Fig. 11. The purpose of the latter is to show that the correct harmonic has been chosen. The single turn loop and bulb is useful to indicate the presence of R.F. in a coil, irrespective of its frequency, and in the case of low power stages will give a rough indication of the amount of such R.F. For example, a 6 V. 0·3 A. bulb fully lit to normal brilliance will indicate that there is about 1·5 to 2 watts of R.F. power in the circuit. That is provided the turn is held close to the coil in question. Obviously in a stage where 20 or 30 watts are expected the loop should be held some distance away and brought very slowly nearer until it lights.

Coil Table

Band Mc/s.	Turns	Approximate dial reading	Link turns
1·8	57	90	4
3·5	22	90	3
7		30	
14	6	60	2

The 1·8 Mc/s. coil is wound with 24 S.W.G. enamelled wire and the others with 22 S.W.G. enamel.

TWO VALVE TOP-BAND TRANSMITTER

The equipment described in this section has been designed to meet a variety of requirements. Essentially, it is a completely self-contained low power transmitter, very suitable for construction by the newly-licensed amateur. As experience is gained, the latter will wish to expand his station and this unit can then become an exciter stage. Incidentally, there are many points in favour of using an exciter which possesses its own power supply.

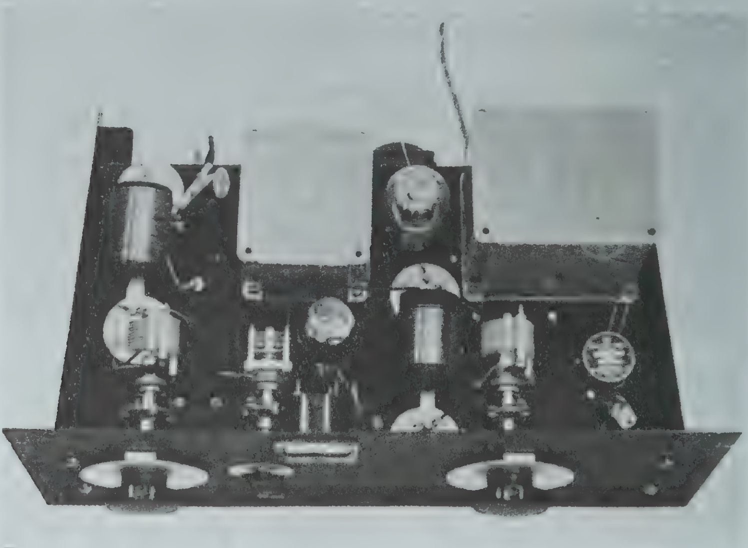
Whilst intended primarily for use within the 1·8 Mc/s. amateur band, it is equally efficient on 3·5 Mc/s. and fulfils all the requirements called for, such as break-in and high reliability. Operation is also possible in the 7 Mc/s. band, at somewhat reduced efficiency.

Several novel features have been incorporated, partly with a view to simplification and partly to enable it to be used for other purposes. These features are the use of (*a*) a Pierce oscillator circuit, which eliminates one tuning control, (*b*) self-bias throughout, (*c*) capacitative coupling between output and aerial circuits, (*d*) junction block connection to power supplies.

The transmitter is flexible as regards the valves that may be used in it—a number of types have been tried, the best combination being a 6V6 as crystal oscillator, followed by a *Mullard* EL37 as amplifier. It is recommended that the 6V6 be retained as C.O., but any beam power tetrode will be found to give good results as an amplifier.

The Circuit

The complete circuit diagram is given in Fig. 12. The first valve is connected as a Pierce oscillator, in which the crystal itself forms the tuned



The two valve top-band transmitter viewed from above.

circuit and no tuning coil or condenser is necessary. The crystal is placed between grid and anode, with a condenser in series to remove the standing D.C. potential from the crystal. A small resistor, R3, soldered directly to the valveholder pin, prevents parasitic oscillation. Bias is secured by a combination of cathode and grid leak resistors and this valve is at all times much under-run. R4 in the anode circuit serves as both decoupling and voltage dropping resistor.

Capacitative coupling to the power amplifier is used. Again, a combination of resistors provides grid bias and it is not possible, under normal circumstances, to over-run the valve. An anti-parasitic resistance is included in the anode circuit and both screen grid and anode circuits are effectively decoupled, the former by R8 and C8, the latter by RFC2 and C11.

It would be possible to connect the aerial to a tap on the P.A. anode circuit, but this practice is not advisable, as it tends to make tuning somewhat more

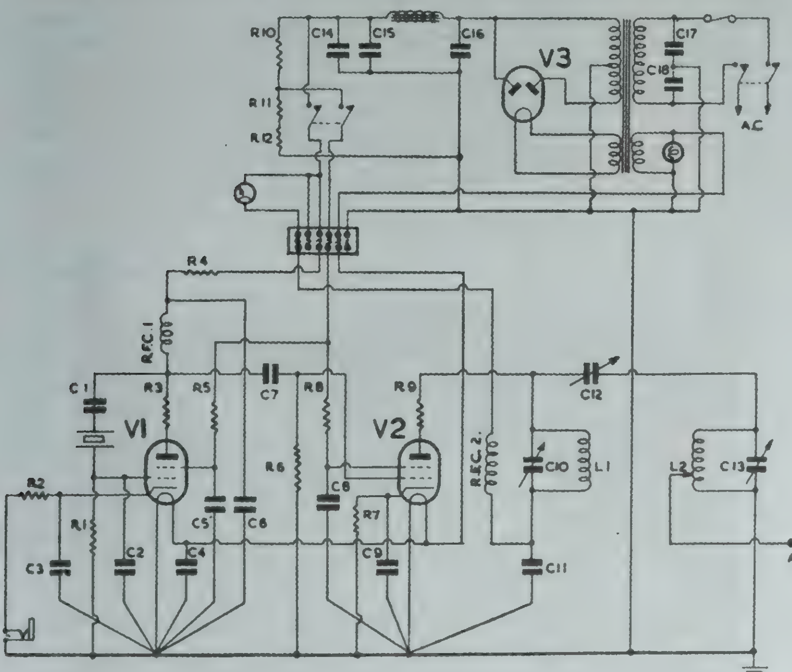


Fig. 12.

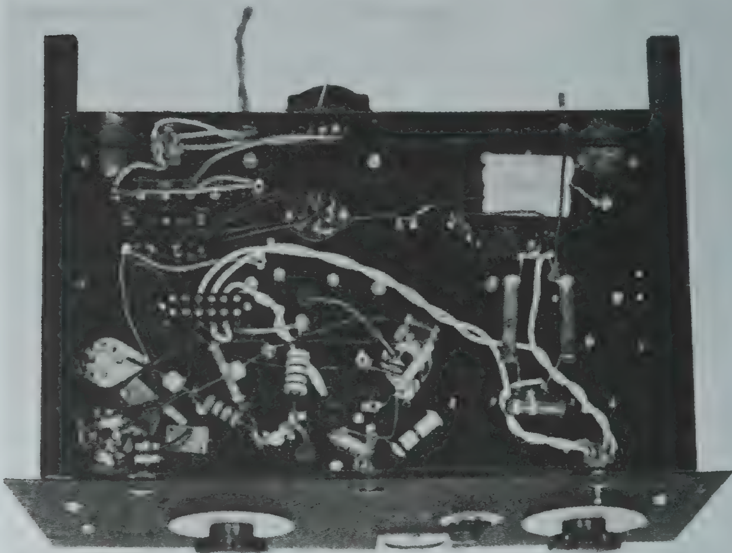
Circuit diagram of the Two Valve Top-Band Transmitter.

C1, 7	200 $\mu\mu$F.	C15, 16	8 μF.	R6	20,000 ohms.
C2	40 $\mu\mu$F.	C17, 18	·002 μF.	R7	400 ohms.
C3, 4, 5, 6, 8, 9, 11	·002 μF.	R1	30,000 ohms.	R8	100 ohms.
C10, 13	160 $\mu\mu$F. variable.	R2	200 ohms.	R9	6 ohms.
C12	15 $\mu\mu$F. variable.	R3	20 ohms.	R10 12	7,500 ohms.
C14	·01 μF.	R4	2,000 ohms.	R11	1,400 ohms.
		R5	1,000 ohms.		

difficult; furthermore it may cause interference to broadcast receivers. A separate aerial circuit should always be used but difficulties crop up in coupling between the aerial and output circuits. Link coupling is good, but it complicates the mechanical design of the coils, unless it is intended to work on one frequency band only, as the number of turns in the link winding will vary from band to band. The somewhat unusual method of top-end capacitative coupling has therefore been employed. This enables the coil design and mounting to be simplified in that one winding only is required in both tank and aerial coils whilst the degree of coupling is easily adjusted. The system works well and no disadvantages have become apparent.

The power supply circuit should be studied carefully. When using a combined H.T. and L.T. transformer, the normal method of switching off H.T. during stand-by periods is to break the H.T. winding centre tap to chassis. This involves the sudden application of H.T. to the smoothing condensers, and is probably the cause of many a breakdown of these components. It is better to leave the voltage across these condensers, but with bleeder resistors connected so that no high voltage builds up (as is possible with a condenser input filter). The bleeder resistors are also used as a potentiometer to supply a reduced voltage to the screen grids of the valves. Note that switching off the H.T. to the anodes of the valves only, is likely to damage them. A double-pole switch is therefore required to disconnect H.T. from both anodes and screens simultaneously, but still leaving the bleeder resistors in circuit.

When carrying out experimental work, it often happens that a readily



Underside View of Two Valve Top-Band Transmitter.

accessible power supply is called for. Mainly for this reason, the connections *from* the power supply components have been wired into one side of a *Grelco* 6-way block, the wiring *to* the valves being taken from the other side of the block. These leads are therefore easily detachable and loose wires can be taken to the experimental apparatus from the block. The meter in the anode circuit of the P.A. valve may or may not be included at will in the external circuit, depending on whether the loose lead is screwed into outlet 5 or 6.

Radio frequency current travelling back through the mains supply is a common but often unsuspected cause of interference to broadcast receivers. To prevent it, in this instance, condensers C17 and C18 are connected across the primary of the mains transformer. These should be rated for at least 500 volts working. The primary circuit also includes a 1 ampere safety fuse and a double-pole switch which, when off, completely disconnects the A.C. supply from the equipment.

The transformer used in the transmitter (illustrated on page 24) gives a smoothed H.T. output of 300 volts, which is about right for operation of the transmitter at an input of 10 watts in the 1·8 Mc/s. band. If it is intended to operate on higher frequencies, where a greater input is permitted, the transformer should be chosen to give up to 400 volts smoothed H.T. In the latter case, the value of R4 will require increasing so that 250 volts on the anode of the first valve is not exceeded.

The bleeder resistance values have been chosen to give a screen voltage of 150, which is suitable for the majority of beam tetrode valves. In the transmitter illustrated, R11 and R12 have been used as a matter of convenience but they may well be combined into a single resistor of 10,000 ohms, rated at 5 watts.

Construction

The chassis, front panel ($8\frac{3}{4}$ in. deep) and side brackets are of mild steel, of international standard size, the whole forming a strong, rigid assembly. The upper chassis view of the transmitter (page 24) shows the layout adopted, the crystal oscillator stage being on the extreme right, the P.A. stage in the middle and the aerial tuning circuit on the left. The mounting holes for the adjustable brackets are drilled $2\frac{1}{4}$ in. from the front of the chassis, which distance will ensure that the slow motion dials fitted to the tuning condensers C10 and C13 and the flexible couplers fit correctly to the condenser spindles, C10 is mounted $4\frac{1}{2}$ in. and C13 mounted $2\frac{1}{2}$ in. from the edges of the chassis. The aerial coupling condenser C12 is fitted $3\frac{1}{2}$ in. to the right of C13. The positions of the other components are by no means critical and may be judged from the photographs. It should be noted that the stand-off insulators holding the coils are mounted $4\frac{1}{2}$ in. apart and are so placed that the wing nuts are easily accessible for rapid coil changing.

All necessary holes for mounting the mains transformer and choke should be made when other holes are drilled but the two components themselves should not be fixed until the wiring has been nearly completed, so allowing the chassis to be handled with greater ease.

The positions occupied by the various condensers and resistors below the chassis can be seen in the photograph on page 26. The mica condensers are bolted to the chassis through the holes provided in the moulded cases. Tag strips are provided where necessary to hold resistors, R.F. chokes, etc., the valveholder tags also serving the same purpose. The 6-way *Grelco* block previously mentioned is fitted conveniently near the mains transformer.

Holes, $\frac{3}{8}$ in. diameter, to take rubber grommets are required beneath the tank coil stand-off insulators, since leads to anode and H.T. pass through the

chassis at these points. The same applies to the leads for the meter. The mains transformer and choke are of the sub-chassis wiring type. If leads have to be brought through from top connections further holes and rubber grommets will be called for, at least for those leads carrying high A.C. and D.C. potentials.

At the rear of the chassis are fitted the mains on/off switch, the fuse holder and an earth terminal. The mains leads are brought out through a ⅜ in. hole fitted with a rubber grommet. On the front are the keying jack (of the open circuit type) standby switch and panel indicating light. These three items are bolted to the chassis proper and ⅜ in. clearance holes, in appropriate positions, should be made in the front panel.

The slow motion dials require holes 1⁵⁄₁₆ in. in diameter, 4 in. from the top of the panel. When marking out, care should be taken that their positions are accurate in the horizontal plane—errors in the vertical plane can be taken up by the adjustable brackets. A small knob controls the coupling condenser through an insulated coupler—a ⅜ in. hole is necessary in the panel to take a bush.

The appearance of the transmitter would possibly be improved if the meter was mounted centrally, but in such a position it would interfere with coil changing. The meter is therefore offset and is located 8 in. from the left hand edge of the panel (not chassis).

A miniature stand-off insulator is fitted at the rear of the chassis, near the aerial coil, and acts as an aerial terminal.

Wiring

The circuit diagram indicates that the earthy sides of all the fixed condensers associated with the crystal oscillator valve V1 have been returned to one single point on the chassis. This feature is important and is one which it is well to follow when constructing any piece of radio equipment. It prevents circulating R.F. currents in the chassis and, by removing the cause of undesirable coupling effects, increases the overall electrical stability.

A stout soldering tag (or several small ones) should be fitted beneath the valveholder fixing bolt remote from the edge of the chassis and to this tag separate wires are soldered from pins 1 and 2 of the valve holder and from condensers C2, C3, C4, C5 and C6.

The same principle is adopted in the case of the second valve. C8, C9 and C11 (one bolt serves to fix all three condensers to the chassis) and pins 1 and 2 of the valveholder are all connected to an earth tag fitted beneath a fixing screw.

The low potential side of the aerial tuning combination L2, C13 is earthed, at one common point only, to a tag fitted beneath a 4BA bolt. From the underside of this bolt a wire is run direct to the earth terminal at the rear.

Before fixing the tank coil stand-off insulators, loops of flexible wire should be fitted beneath the bolt heads inside the hollow portion and led through the chassis. The lead nearer the panel is soldered to a tag strip and thence connects *via* the R.F. choke to outlet 6 in the *Grelco* block. The other lead goes *via* R9 direct to tag 3. Both R9 and R3 are soldered directly to the valveholder tags with very short wires.

The remainder of the wiring is straightforward. Heavy gauge wire, either solid or flexible should be employed for the heater and R.F. tuned circuits, 20 S.W.G. tinned wire enclosed in insulating sleeving serving elsewhere. Good quality flex is advisable for the meter and mains leads.

When all other wiring has been completed, the mains transformer and choke are bolted into position and wired in.

Coils

The coils, both tank and aerial, for use on the 1·8 Mc/s. band consist of 50 turns 20 S.W.G. enamelled wire on paxolin formers, 2¾ in. diameter. Strips of brass, approximately 1¼ in. long by ½ in. wide, are bolted to the ends of the coil and holes drilled so that they fit over the 2BA bolts in the stand-off insulators. Four tappings at the earthy end are wanted on the aerial coil to accommodate different lengths of aerial. The first tapping is 4 turns from the end, the others each 4 turns further away.

On 3·5 Mc/s. 24 turns are required, and, as the formers will then take a heavier gauge of wire, 16 or 18 S.W.G. enamelled copper wire should be used. The tapping points in this case should each be three turns apart.

The coils for 7 Mc/s. operation should be either of the self-supporting type or wound on a ribbed former. The number of turns required is twelve assuming a diameter of 2½ in. or ten if of 3 in. diameter. The aerial coil should be arranged so that it is possible to tap on to each turn separately.

Operation

After connecting up the mains supply, ensure that the heaters of the valves are receiving the correct voltage (6·3 or near). Closing the stand-by switch will apply H.T. to the valves and, with the cathode circuit of V1 open, the meter in the anode circuit of V2 should indicate a current of approximately 30 mA. This continual drain will do no harm—in fact it will improve the regulation of the power supply.

A meter plugged into the keying jack should show about 20 mA. and, at the same time, the anode current of V2 should rise to 50 mA. or more.

No time should be lost in tuning C10 to the point where the anode current drops to a very low value. A small dip may be noticed at a low dial reading, at which point the circuit is tuned to the second harmonic. Further rotation will result in a pronounced dip. During this process, C12 should be set at minimum capacity and C13 at maximum.

The transmitter is then ready for use. Connect a good low resistance earth to the earth terminal and the aerial to the miniature stand-off insulator. From the latter, a short flexible lead terminating in a crocodile clip is used to tap on to the aerial coil. The tap will vary according to the length of aerial which may be anything from 30 to 200 feet—the higher the better. A long-wire aerial should be tapped near the *earthy* end of the coil, a short-wire aerial well up the coil.

On 1·8 Mc/s. the coupling condenser will be set at about one-third capacity—4 on the scale. Very rarely will more coupling be required and on the higher frequencies it will always be less. For instance, on 7 Mc/s. the coupling condenser should be set at zero, so that only the normal minimum capacity is effective.

On rotating C13 to resonance the anode current of V2 will rise. If the rise and fall is sharp, the aerial is not matching-in properly and the tapping point should be increased. Conversely, if the rise and fall is sluggish, and of comparatively small amplitude, the tapping point should be reduced. The aim is to secure a smooth rise and fall, with the optimum meter reading such that the required input is obtained. This will be about 30 mA., on 1·8 Mc/s. and up to 50 mA., possibly at a greater H.T. voltage, on other frequencies.

Keying

Keying is effected in the cathode circuit of V1. This valve is completely cut-off when the key is up and no harm will result if the transmitter is left

switched on, with H.T. applied, for long periods. Keying of the original transmitter is smooth, but if any clicks become apparent, a condenser of 0·1 μF. capacity, 1,000 volts working and a resistor of 2,000 ohms may be connected in series across the key. For operation by remote control, a relay can be connected in place of the key.

The transmitter, as shown, is not suitable for telephony operation without small modifications. Chiefly, these consist of reducing the values of C8 and C11 and of altering the system of feeding the screen of V2 to enable simultaneous screen and anode modulation of this valve.

Minor Modifications

In some instances the constructor may wish to vary the details somewhat. Slow motion dials are a definite help when tuning the transmitter, but direct drive dials may be substituted.

The moving coil milliammeter, which is shown permanently wired to the anode circuit of the P.A. valve, is very necessary when tuning-up but, of course, an external meter may be employed. If desired, an insulated closed circuit jack may be fitted at either the front or rear of the chassis for this purpose.

If the transmitter is required mainly for use on 7 Mc/s. the values of C10 and C13 should be changed to 60 $\mu\mu$F. as this will render tuning easier.

When experience of tuning the transmitter has been gained, the value of the cathode resistor R7 may be reduced to 200 or 250 ohms. A slight increase of both input and output is probable but the optimum value of resistance is bound-up with the type of valve employed in the P.A. stage.

WWV Standard Frequency Transmissions

Frequency Mc/s.	*Modulation c/s.*	*Frequency Mc/s.*	*Modulation c/s.*
2·5	440	20*	440 & 4,000
5	440	25	440 & 4,000
10*	440 & 4,000	30	440 & 4,000
15*	440 & 4,000	35	440 & 4,000

In the United Kingdom transmissions marked (*) are received more reliably than others, although signals of considerable strength are obtainable on 25 and 30 Mc/s. when propagational conditions are favourable. The carrier frequencies are accurate to at least one part in 50 million.

CHAPTER 4 FIVE BAND 25 WATT TRANSMITTER

ALTHOUGH intended primarily for the holder of a 25-watt licence, this is a piece of apparatus for which use might be found in almost any station, either as a stand-by transmitter, or as a drive unit for a high power final amplifier.

Facilities

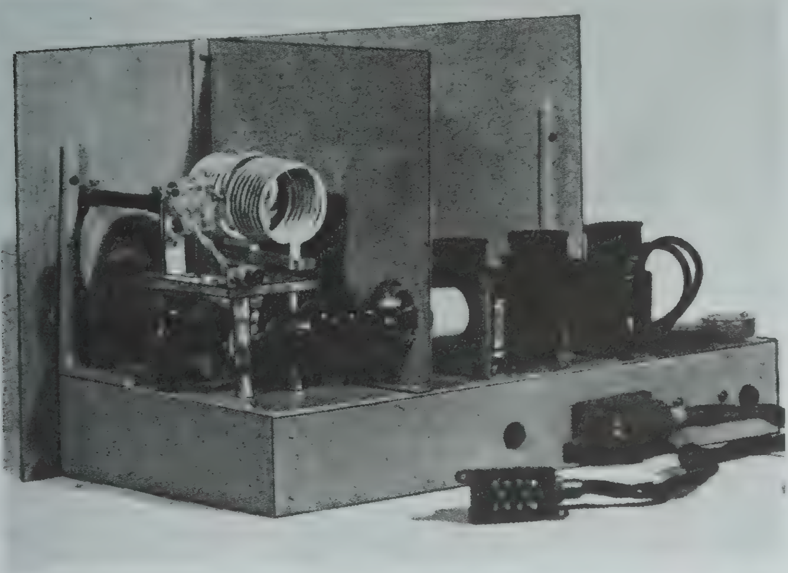
A full 25 watts input may be employed on C.W. on the 3·5, 7, 14, 21* and 28 Mc/s. amateur bands using only one crystal, and with the final valve operating as a P.A. in each case. This is made possible by the incorporation of modern high-efficiency British valves in all stages.

Two switched meters are provided for checking operating conditions at appropriate points, in addition either of two crystals may be brought into service at the touch of a switch. Provision is made for break-in working and for master-oscillator control if desired. A separate power unit supplies all anode, heater and grid voltages for the transmitter. Although no provision is made in the design for modulating the P.A., this may be achieved, with the addition of a suitable modulator and power supply, as described later in this chapter.

The Circuit.

It will be seen from the theoretical diagram, Fig. 13, that a three-valve

* Not yet released for amateur service.



Rear view of the 25 watt transmitter. The sockets for the P.A. tank padding condenser can be seen just below the polystrene base of the tank unit.

circuit is employed, *Mullard* QVO4–7 tetrodes being used in the first two positions with a *Standard Telephones* type 5B 250A—the equivalent of the well-known American type 807—as the P.A.

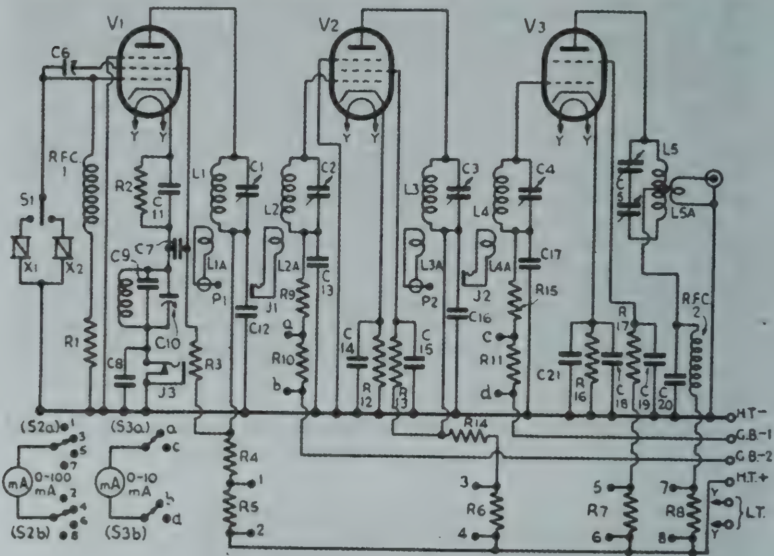


Fig. 13.
Circuit diagram of 5-band 25 watt transmitter.

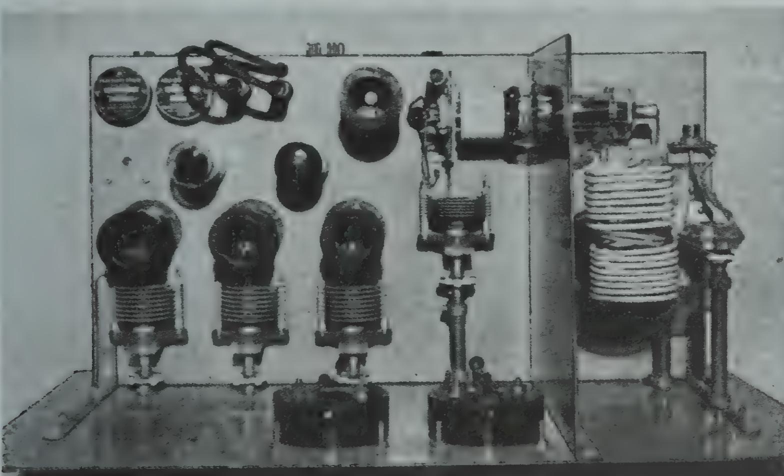
C1, 2, 3	60 $\mu\mu$F. *Eddystone* No. 1093.	R2, 12	500 ohms. *Erie* type 3, 3 watt.
C4	100 $\mu\mu$F. *Eddystone* No. 1130.	R3, 13, 17	30,000 ohms. *Erie* type 16, 5 watts.
C5	65 + 65 $\mu\mu$F. *Q-Max* P.A. tank unit.	R4	8,000 ohms. *Erie* type 16, 5 watts.
		R5, 6, 7, 8	100 ohms. *Erie* type 8, ½ watt.
C6, 9	3–30 $\mu\mu$F. *Mullard* air trimmer.	R9	20,000 ohms. *Erie* type 8, ½ watt.
C7, 8	·01 μF. *T.C.C.* type CP45W.	R10, 11	200 ohms. *Erie* type 8, ½ watt.
C10	60 $\mu\mu$F. *Eddystone* Nos. 581 or 582.	R14	8,600 ohms. *Erie* type 16, 5 watts.
C11, 12, 13, 14, 15, 16, 17, 18, 19	·005 μF. *T.C.C.* type CP45W.	R15	10,000 ohms. *Erie* type 8, 1 watt.
		R16	200 ohms. *Erie* type 0, 3 watts.
		RFC1	*Eddystone* No. 1022.
C20	·001 μF. *T.C.C.* type CP45W	RFC 2	*Eddystone* No. 1010.
C21	2 μF. *T.C.C.* type CE30G.	V1, 2	*Mullard* No. QVO4–7.
R1	50,000 ohms. *Erie* type 8, ½ watt.	V3	*S.T.C.* No. 5B/250A.

The coil across C9 is an R.F. choke similar to R.F.C.1.

The condenser C6 is only required with sluggish crystals and should then be adjusted to approx. ·10 $\mu\mu$F.

Valve-holders	B9G Ceramic *Webb's Radio.*
Meter Switches	2-bank 3-way/*Webb's Radio.*
	2-bank 7-way/*Webb's Radio.*
P.A. Tank Unit	*Q-Max*, Mk. I and all coils.
Output co-axial plug and socket ...	L619/P and L604/S. *Belling-Lee.*
Meters	0–10 mA. and 0–100 mA. *Pullin* type 25 square flush.
Power Plugs and Sockets	10-way. *Eddystone* type 534 and 535.
Knobs (Switches)	*Eddystone* type 1044.
Dials	*Eddystone* type 595.
Extension Controls	*Eddystone* type 1008.
Coil Bases	*Eddystone* type 964.
Flexible couplers	*Eddystone* type 529.
Chassis, screen and panel	*Philpott's.*

By using a regenerative crystal oscillator circuit, tuning at crystal frequency is avoided, so eliminating one control. At the same time V1 is able to act as an efficient frequency multiplier when required. Many amateurs look askance at regenerative crystal oscillator circuits, regarding them as unreliable, difficult to tune and liable to damage crystals due to excessive crystal current brought about by careless handling of the controls. Such criticisms are well-founded in regard to *some* regenerative circuits, but that employed in this transmitter, although not so well known as, for instance, the almost universal *Tritet*, is much simpler to handle and has the advantage that the maximum crystal current (around 30 mA.) actually *decreases* when the anode circuit is tuned to the fundamental frequency of the crystal, or to the second or third harmonics. This valuable feature makes for simplicity in operation as no harm can possibly be caused to a crystal while tuning-up on any frequency.



Plan view of the transmitter. The S.T.C. 5B/250A P.A. valve can be seen protruding through the vertical screen. The co-axial output socket is to right of the 14 Mc/s. coil in the tank unit.

The output from V1 is taken from a link winding on the 6-pin *Eddystone* coil former L1 through a short length of flex to a telephone-type plug (P1) which may be inserted into either one of two jack sockets J1 and J2. The socket J1 is connected to a link winding on the grid coil (L2) of V2, and is employed when additional frequency multiplication of the output of V1 is required. J2 connects in a similar manner to the grid coil of the P.A. stage (L4). In the latter position the output from V1, and therefore the input to V3, could be at the fundamental frequency of the crystal or its second harmonic. J1 and J2 also serve as convenient input points for an external V.F.O. if such is in use.

Although telephone-type plugs and jacks are not constructed with insulating material which possesses particularly good R.F. properties, it must be borne in mind that the R.F. voltages involved in a link line are quite low, and no appreciable losses, even at 28 Mc/s., have been traced to the use of these convenient components.

Keying is carried out in the cathode circuit of V1, so that break-in working is possible, the anode current of the other stages being held at a safe figure by the standing grid bias. If C8 is not found to reduce key clicks to a low enough level, any of the standard keying filters may be incorporated.

The Frequency-Multiplying Stage.

The *Mullard* QVO4-7 is well suited for use in this position as its requirements in respect of grid drive are very modest. This ensures satisfactory operating conditions even when the input from the previous stage is on one of the harmonics of the crystal in use. L1 and L2 must, however, always be tuned to a similar frequency in order that energy may be transferred, *via* the link windings, between the two circuits. A frequency multiplication of 2, 3, or 4 times is easily possible in V2, although under normal conditions only the even harmonics are used.

Plug-in coils, on 6-pin *Eddystone* formers, are employed in the first four positions, and full details of their windings will be found in the coil data section.

The Power Amplifier.

The *S.T.C.* type 5B/250A tetrode has been chosen for this position for several reasons: its drive requirements are modest—4 mA. grid current suffices in most cases—it operates with good efficiency on all the bands covered by the transmitter, and being a standard type, equivalent valves are produced by several manufacturers.

It has already been mentioned that the grid circuit inductance is a plug-in type similar to those employed in the earlier stages, while the anode tuned circuit consists of a *Q Max* tank unit comprising a series-gap condenser, sockets for centre-tapped plug-in coils and a swinging-link with coupling control brought out to the front panel. As supplied the 65 + 65 μμF. tuning condenser has a wiping contact to the rotor spindle brought out to a tag on the right-hand side of the assembly. As this contact is not required in this circuit, the wiper is removed and the tag utilised as an earthing point for the anode bypass condenser (C20), which is fitted under the polystyrene base of the unit immediately below the coil sockets.

Three coils are required to cover the 7, 14 and 28 Mc/s. bands, and these may be obtained from *Q Max*. For 3·5 Mc/s. operation the 7 Mc/s. coil is used, loaded with a fixed condenser of *U.I.C.* manufacture, type C 117, CSW.B specially designed for such applications where a large R.F. circulating current has to be passed. It is important that the condenser specified, or one with similar R.F. characteristics be used. Ordinary mica dielectric types are *not* suitable for this purpose. The capacity of 65 μμF. results in an optimum ratio of inductance to capacity being obtained with this valve when operating on the 3·5 Mc/s. band.

For ease in band-changing this loading condenser is mounted on a plug consisting of a piece of polystyrene 2 in. by ½ in. by ¼ in. thick in which are mounted two pins set at 1⅜ in. centres which fit into sockets soldered one to each of the two banks of stator vanes in the tuning condenser horizontally on the right-hand side when viewed from the front.

Out put from the P.A. is taken *via* two twisted, flexible wires from the moveable link coil plug to a *Belling-Lee* co-axial cable socket mounted on a small brass angle-bracket secured by one of the tank unit fixing bolts.

Metering.

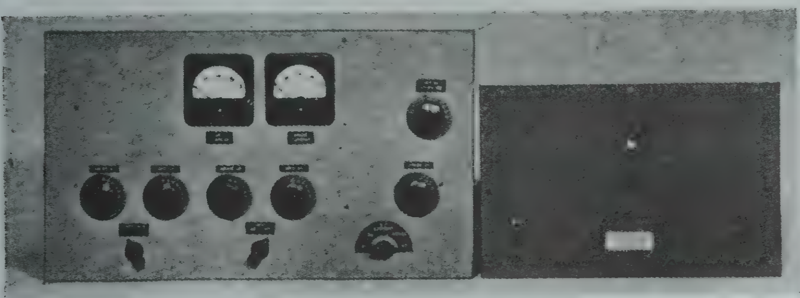
Two *Pullin* meters are provided on the front panel, that on the left reading grid current, in either V2 or V3 according to the position of the switch, while

the other, with the aid of a four-position switch reads the anode currents of V1 and V2 and the screen grid and anode currents of the P.A. (V3). In the case of both these switches connections are made only to alternative contacts so that a blank space occurs between each to guard against possible short circuits.

Construction.

The transmitter is built on an aluminium chassis $\frac{1}{16}$ in. thick measuring $16\frac{1}{2}$ in. by 9 in. by $2\frac{1}{2}$ in. deep, with a panel $18\frac{1}{2}$ in. by 11 in. A screen of the same material, set $11\frac{3}{4}$ in. from the left-hand end of the chassis, divides the P.A. from the remainder of the circuit. The P.A. valve protrudes through this screen as will be seen in the photograph, its centre line being $1\frac{3}{4}$ in. above the chassis.

The four tuning condensers are mounted on small polystyrene brackets just clear of the chassis, with the first three coils immediately to their rear. The fourth coil, *i.e.* the one belonging to the grid circuit of the P.A., is



Front view of the 25 watt transmitter and power pack.

mounted to the left of the P.A. valve-holder, its tuning condenser being slightly off-set, and operated from its panel control *via* an *Eddystone* extension rod. The valves V1 and V2 are set two inches from the rear of the line of coils, and have the same mutual separation as the latter—and the four tuning condensers—of $2\frac{3}{4}$ in. In order to avoid possible strain on the condensers due to their shafts being slightly out of line with the dials these condensers are driven through *Eddystone* flexible couplers.

Two *Q.C.C.* type M crystal sockets are mounted $1\frac{1}{2}$ in. centre-to-centre at the left-hand rear edge of the top of the chassis, with the *Wearite* two-way crystal switch S1 on the rear drop of the chassis below, and slightly to the right of them. The operating knob for this switch is brought out to the front of the panel, in line with the two meter switches, by means of an *Eddystone* extension rod.

The regeneration condenser, C10, occupies a position in line with, and to the left of V1. The component in this transmitter is an *Eddystone* type 581 made for screwdriver adjustment, and a hole was therefore cut in the chassis above it for this purpose. As this type of condenser has now been discontinued, it is recommended that a type 582 be substituted. As the 582 is provided with a normal shaft and one-hole fixing bush it should be mounted by this means, and a knob provided for adjustment.

Tag boards are employed to support certain of the larger resistors, but all

bypass condensers, and most of the smaller resistors, etc., are held firmly in the run of the wiring. Bypass condensers are taken from the appropriate tag on the valve or coil sockets directly to an earth point on the chassis by the shortest route. Care should be exercised to ensure that the earthing tags on the centre spigot socket of the two B9G valve-holders are connected to the chassis.

As there are a fair number of controls on the panel, these have been provided with labels indicating their function. The labels, apart from their utilitarian aspect, make a worthwhile contribution to the appearance of the finished instrument, and may be obtained from *T. A. Butler & Co.* (1927) *Ltd.*, 48–52, Vittoria Street, Birmingham, 1.

Operation.

On insertion of a telephone plug into the jack J3, mounted below the crystal holders on the rear apron of the chassis, the transmitter should key satisfactorily, C1 being adjusted to the H.F. side of the resonance dip in the usual way, to produce the best note.

A crystal in the range 1·75 to 1·9 Mc/s. may be employed for control in the 3·5 Mc/s. band without any difficulty, the first valve then acting as crystal oscillator and doubler.

Using only two crystals—one with a fundamental on 3·5 Mc/s. and one with a fundamental on 7 Mc/s.—the transmitter may be tuned up in many different ways on the five bands. A table showing the various possibilities is given in page 39. It is again stressed that although in several cases V2 is not in use, coils must *at all times* be inserted in the L2 and L3 sockets in order to maintain circuit continuity. It is immaterial which one of the spare coils is employed for this purpose. The positions of the two link plugs P1 and P2 are also indicated in the table.

While this is by no means a complicated transmitter to handle, it is extremely versatile, and in order that tuning up may be accomplished without difficulty an account of tuning procedure will be given, for output on the 28 Mc/s. band using 3·5 and 7 Mc/s. fundamental crystals.

Insert the two crystals in their respective sockets, and operate the crystal switch so that the 7 Mc/s. crystal is in circuit. Insert 7 Mc/s. coils in positions L1 and L2, and 28 Mc/s. coils in positions L3 and L4 and also in the P.A. tuning unit. The number of turns on the output link coil L6 must be determined by trial, but two turns should be used to start with. A length of co-axial feeder, terminating in a *Belling-Lee* co-axial plug, should be inserted into the socket on the P.A., and the other end connected to the aerial tuning unit in use. In place of the aerial a dummy load should be used during these preliminary tests to avoid causing interference to other users of the band. A 40 or 60 watt electric lamp is suitable for this purpose.

Plug P1 should be inserted into J1, and plug P2 into J2. This will bring V2 into circuit. The regeneration condenser C10 should be set at *full* capacity (minimum feed-back), and the grid meter switch S2 turned to its left-hand position to read the grid current of V2. The anode meter switch S3 should also be turned to its extreme left-hand position for measurement of the anode current in V1.

The heaters having been lit for half a minute or so, H.T. may now be applied by means of the switch S4 on the power unit, and C1 turned to a position where the anode current of V1 is at a minimum. This, with a crystal of about 7·1 Mc/s. will be around 65° on the dial, at which point the anode current of V1 will be approximately 25 mA.

Next resonate L2 by means of C2, which will read about the same as C1,

and the grid current of V2 should be in the region of 2 mA. It is very easy to over-drive V2, so C2 should be detuned so as not materially to exceed this figure. Turn the anode current meter switch to its next position, to read anode current in V2, and quickly bring L3 to resonance with the aid of C3. The dial reading will be about 75° and the current about 25 mA. Check that the grid drive to V2 remains at a figure between 1 and 2 mA., readjusting C2 as necessary.

The grid meter switch is now turned to its right hand position to read grid current in V3, and C4 tuned for the highest reading, which should be between about 2·5 and 4 mA. Move the anode current meter switch to the next but one position clockwise—anode current of the P.A.—and resonate the P.A. anode circuit by means of C5. This operation should be done as quickly as possible, as otherwise the P.A. anode current may be well in excess of the permissible value. Having obtained the lowest reading for the P.A. anode current, tune the aerial coupling circuit, adjusting the coupling between L5 and L5A as necessary, until the P.A. is drawing its rated 25 watts. This should be indicated by the load lamp glowing brightly.

It is essential during the tuning-up operation to check each tuned circuit with a simple harmonic indicator (as described elsewhere in this book). In the case just described this should be L1/L2–7 Mc/s. L3/L4/L5 and the aerial tuning circuit—28 Mc/s. Steps must also be taken to see that all circuits are drawing correct anode and grid currents.

If the frequency of the 3·5 Mc/s. crystal occupying the other crystal socket is fairly close to half that of the 7 Mc/s. crystal, a turn of the crystal switch should enable almost the same output to be obtained, the first valve then acting as a crystal oscillator and doubler. If the output is not satisfactory, even after a complete retune of all circuits, the regeneration condenser, C10, should be decreased in value, but only so far as is necessary to obtain the correct grid current in V2.

The foregoing may seem extremely complicated, but this is only because any such detailed description takes much longer to read than to carry out the operations. The tuning-up procedure should be carried out on each band and the dial readings, etc., recorded for future reference.

It should be borne in mind that when tuning-up a multi-stage transmitter there is always a possibility that the later stages will be over-run whilst earlier circuits are being adjusted. It is, therefore, advisable to tune the various grid circuits so that only the minimum amount of grid current is produced to provide drive to the following stage until the corresponding anode circuit is at resonance. As the P.A. anode current will fall to a very low value at resonance without an R.F. load or aerial being connected, it is imperative that the output valve should not be operated in that condition for longer than is absolutely necessary, to ensure that the anode circuit is in resonance, and then only with reduced grid drive.

Grid Bias Adjustment.

Both V2 and V3 derive their grid bias in three different ways; by rectified R.F. drive across the grid resistances R9 and R15, by voltage drop in the two cathode resistors R12 and R16, and by the standing bias provided by the power pack. The adjustment of the standing bias is described under the section dealing with the power supply.

The 2 μF. condenser C21 is only required if the P.A. is to be modulated, and may be omitted if C.W. operation only is contemplated. For the same reason the P.A. anode bypass condenser C20 does not exceed ·001 μF., but may be increased to at least ·005 μF. if modulation is not to be used.

Modulating the Transmitter.

This transmitter is suitable for telephony operation by anode and screen modulation of the P.A. valve. In order to achieve this a slight modification only is required in the H.T. feed arrangements to the P.A. anode and screen. Both R7 and R8 should be disconnected from the H.T. positive line, connected together, and a lead taken to one side of the secondary winding of the modulation transformer. The remaining side of the transformer secondary is connected to the H.T. positive line. Ample space exists on the rear drop of the chassis for mounting a socket for this purpose which must, of course, be bridged when C.W. operation is required.

Power Supply.

This unit which may be obtained complete from *Radiocraft* provides approximately 500 V. of H.T. at up to 100 mA. with excellent regulation, a 6·3 V. heater supply, and two adjustable negative grid bias voltages for V2 and V3 respectively. The circuit diagram is shown in Fig. 14.

The 100 watt bleeder resistance R18 is mounted under the chassis, and is provided with two adjustable tapping points. One of these is connected to the chassis and forms the H.T. negative supply line. The H.T. current flowing through the portion of the resistor between this point and the end of the resistor to which S4 is connected is arranged to drop approximately 70 V. *when the transmitter is drawing its normal current.* This constitutes the negative grid bias for the P.A.

A further tap, intermediate between the P.A. grid bias connection and the chassis, supplies a lower negative voltage of approximately 45 V. for V2.

The tapping points are determined by the positions of the clips on the resistor, and these may be adjusted by slackening the nut and bolt on the clip

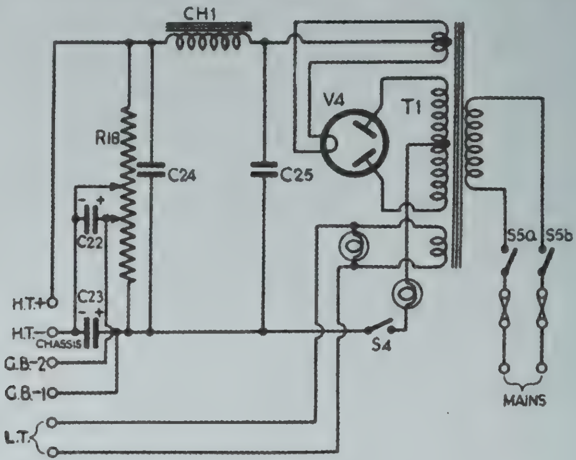


Fig. 14.
Circuit diagram of power pack used for 25 watt 5-band transmitter.

C22, 23	8 μF. Dubilier "Drilitic"	V4	U52 *Osram.*
C24, 25	4 μF.	T1	600–0–600 V. *Radiocraft.*
R18	20,000 ohm, 100 watt.	CH1	Smoothing Choke „

and moving the latter to the required position. Care must be exercised to ensure that the resistance wire is not damaged during this adjustment, and that the clips make firm contact when in position. As a guide to preliminary adjustment the centre of the clip connected to the chassis should be approximately 1 in. from the negative end of the resistor, but its exact position must be determined by measuring the bias voltage at the P.A. by means of a high resistance voltmeter connected between the "earthy" end of L4 and the chassis. About 100 V. will be found to be correct, and of this some 70 V. is derived from the power pack.

For V2 about 45 V. is required from the power pack, but as the operating conditions of this valve are to a large extent self-adjusting, the aim should be

Band Mc/s.	Crystal Mc/s.	L1	L2	L3	L4	L5	P1	P2	REMARKS
3·5	3·5	3·5	—	—	3·5	7	J2	—	65 μμF. condenser in parallel with L5.
7	3·5	7	—	—	7	7	J2	—	
7	7	7	—	—	7	7	J2	—	
14	3·5	14	—	—	14	14	J2	—	
14	3·5	7	7	14	14	14	J1	J2	
14	7	14	—	—	14	14	J2	—	
21	3·5	7	7	21	21	21	J1	J2	28 Mc/s. coils at L3, 4. 14 Mc/s. coil at L5.
21	3·5	10·5	10·5	21	21	21	J1	J2	14 Mc/s. coils at L1, 2, 5. 28 Mc/s. coils at L3, 4.
21	7	7	7	21	21	21	J1	J2	28 Mc/s. coils at L3, 4. 28 Mc/s. coil at L5.
21	7	21	—	—	21	21	J2	—	28 Mc/s. coils at L1, 4, 5.
28	3·5	7	7	28	28	28	J1	J2	
28	3·5	14	14	28	28	28	J1	J2	
28	7	7	7	28	28	28	J1	J2	
28	7	14	14	28	28	28	J1	J2	

Table showing the different ways in which the transmitter may be set up for operation on the various bands. On all bands except 21 Mc/s. the figure shown under L1 L2, etc., indicates the *coil* to be used in that position. For 3rd harmonic operation in the C.O., the figure shown is the *frequency* to which the circuit is tuned; the appropriate range of coil is then given in the Remarks column.

so to balance the grid bias and the output on the highest frequency as to limit the standing current in V2 to not more than 10 mA. *under key-up conditions.*

Coil Data.

Two of each coil required.
All windings are of 26 S.W.G. enamelled wire wound on *Eddystone* type 537 6-pin formers.

3·5 Mc/s.	..	32 Turns winding length 1·7 in.
		3 Turns link winding close to earthy end. A 40 $\mu\mu$F. mica condenser (*T.C.C.* SMW 5%) is mounted inside the coil former and connected across the main winding on this range only.
7 Mc/s.	..	19 Turns winding length 1 in.
		3 Turns link winding close to earthy end.
14 Mc/s.	..	10 Turns winding length ½ in.
		2 Turns link winding spaced one wire diameter from earthy end.
28 Mc/s.	..	4 Turns winding length ½ in.
		1 Turn link winding ¼ in. from earthy end.

CHAPTER 5 A SIMPLE V.F.O. UNIT

The Circuit

IN any V.F.O. unit the chief consideration is one of stability, both in a mechanical and an electrical sense.

There are a number of popular circuits in use at the present time, any one of which will produce a good stable signal; the one chosen for description here is known as the "Clapp Oscillator." This arrangement possesses the

Front view of variable Frequency Oscillator.

advantage of simplicity and reliability, and although the output is relatively small it is sufficient to drive a modern small tetrode valve.

From an examination of the circuit shown in Fig. 15 it will be seen that the coil L is series-tuned by the condenser C1, which is in parallel with the fixed condenser C2. The condensers C3 and C4 must be of good quality, and the junction of the two provides the cathode tap of the oscillator circuit.

The use of a 6SN7 double-triode enables complete isolation to be obtained from the later stages of the transmitter to which the unit will be coupled, thereby helping to ensure that degree of stability which is so essential in any V.F.O. In addition, a VR150 30 voltage stabiliser is used to regulate the H.T. supply to the oscillator section of the valve. This considerably reduces the possibility of chirp when the oscillator is keyed.

Output is taken from the cathode of the oscillator section of the valve to the grid of the second half of the 6SN7, and the final output of the unit is

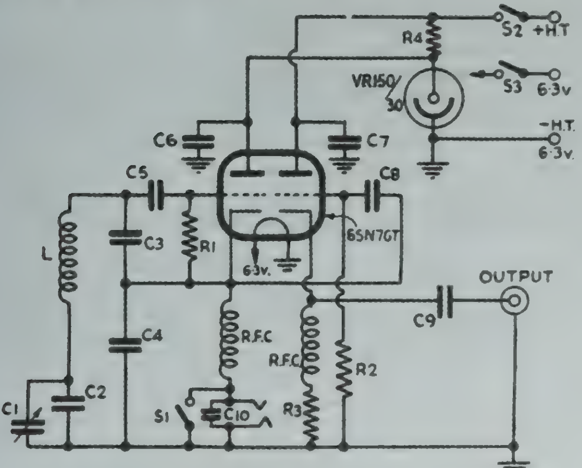


Fig. 15.
Circuit of Variable Frequency Oscillator.

L	62 turns 22 S.W.G. enamelled wire on 1½ in. former. *Eddystone*, type 537.
C1	60 $\mu\mu$F. max. variable condenser. *Eddystone*, type 582.
C2	75 $\mu\mu$F. fixed condenser. *T.C.C.*, type .
C3, 4	·001 μF. fixed condenser. *T.C.C.*, type CM24N 10 per cent.
C5, 8	100 $\mu\mu$F. fixed condenser. *T.C.C.*, type CC31y.
C6, 7, 10	·01 μF. fixed condenser. *T.C.C.*, type 543.
C9	100 $\mu\mu$F. fixed condenser. *T.C.C.*, type CC42h.
R1	25,000 ohms. *Erie*, type 8.
R2	25,000 ohms. *Erie*, type 8.
R3	1,000 ohms. *Erie*, type 8.
R4	5,000 ohms. *Erie*, type 0.
R.F.C.	*Eddystone* R.F. choke, type 1010.

Other components:
- Chassis. *Eddystone*, type 643.
- Cabinet. *Eddystone*, type 644.
- Slow motion drive. *Eddystone*, type 637.
- Flexible coupler. *Eddystone*, type 529.
- 2 Octal valveholders. *McMurdo*, type POP8/US.
- 1 6SN7GT valve. *Brimar*.
- 1 VR 150/30 valve. *Brimar*.
- 3 on-off switches. *Bulgin*, type S261.
- 1 Panel lamp. *Bulgin*, type D50.
- 1 3-pin socket. *Belling Lee*, type L1113.
- 1 Concentric socket. *Belling Lee*, type L604/S.
- Panel name plates. *T. A. Butler & Co.*

taken from the second cathode circuit. This gives a low-impedance output which may be coupled to the transmitter *via* a length of 80-ohm concentric cable. The V.F.O. can thus be placed on the operating table several feet away from the transmitter.

If a greater output is required a tuned circuit may be included in the anode circuit of the second half of the 6SN7, and the output taken from this coil *via* a two turn link. The output of the unit as shown, will, however, usually prove sufficient.

As with all equipment described in this book, the design of the V.F.O. has been kept as simple and straightforward as possible, consistent with stability.

Power Supply

The power supplies for the unit are not included on the same chassis, for apart from cramping the layout, the additional heat produced would tend to increase the amount of frequency drift of the V.F.O. A supply of 250 volts D.C. and 6·3 volts A.C. is required.

Since the frequency of operation must be known at all times a standard-frequency crystal calibrator is necessary. Such a unit should always be available for simultaneous use with the V.F.O. A suitable calibrator is described later, the design of which permits the power supply for the V.F.O. to be obtained from the built-in power pack of the frequency meter.

Construction

The V.F.O. unit has been designed to fit an *Eddystone* die-cast chassis, type 643, and the whole unit can be enclosed in a small *Eddystone* cabinet, type 644.

The greatest care must be taken with the mounting of the components to ensure absolute mechanical rigidity. This applies especially to all components associated with the inductance L. The variable condenser C1 is mounted on the top of the chassis on a stout bracket made from ⅛ in. brass, while the condensers C2, 3 and 4 are mounted below the chassis on *Bulgin* tag-strips. The layout of the underside of the chassis is shown in Fig. 16 from which the various components can be readily identified.

The inductance L is wound on an *Eddystone* former, type 537, from which the pins have been removed. This enables the coil to be bolted firmly to the chassis by means of a 2BA bolt and washer. The leads from the coil are taken through the chassis *via* the holes left by the pins.

A winding of 62 turns of No. 22 S.W.G. enamelled wire is wound on to the former as tightly as possible with the turns just touching. Before commencing winding two small holes should be drilled in the former, 2 in. apart so that the ends of the wire may be passed through them and thence down through the holes in the base where the pins were originally fitted. The coil may be doped with " Denfix," or other good-quality polystyrene cement. Great care must be taken to ensure that no turns are loose; particular attention being paid to the end turns.

With components of the values specified it will be found that the unit will tune over the 1·7 to 2 Mc/s. band comfortably, with a little to spare at either end. The output may then be taken to the transmitter proper, and the frequency multiplied, if desired, in the manner described earlier in this booklet.

As will be seen from Fig. 15, keying is carried out in the cathode of the oscillator section of the valve, the connections for this being brought out to an *Igranic* jack mounted on the rear edge of the chassis, together with the power

supply socket, which is a 3-pin *Belling Lee*, type L1113. The output concentric socket, also *Belling Lee*, type L604/S, is mounted in the centre.

A switch, S1 in Fig. 15, is shown connected across the keying jack, and is mounted on the front edge of the chassis, together with S2 and S3, the H.T. and L.T. switches respectively. A panel indicator lamp is also mounted on the front edge, and is wired in parallel with the L.T. supply. The purpose of S1 is to enable the unit to be conveniently switched on for adjustment of frequency, testing or for telephony operation.

In order to secure the chassis firmly to the cabinet, two long 2BA bolts are required. Suitable holes for these are provided in the chassis and cabinet. They may be made from lengths of 2BA screwed rod with a nut at each end. In addition the switches should project through the front panel and be clamped by extra nuts from the outside. These extra nuts can be obtained from the makers of the switches.

Name plates, which add an attractive finish, can also be fitted to the front panel.

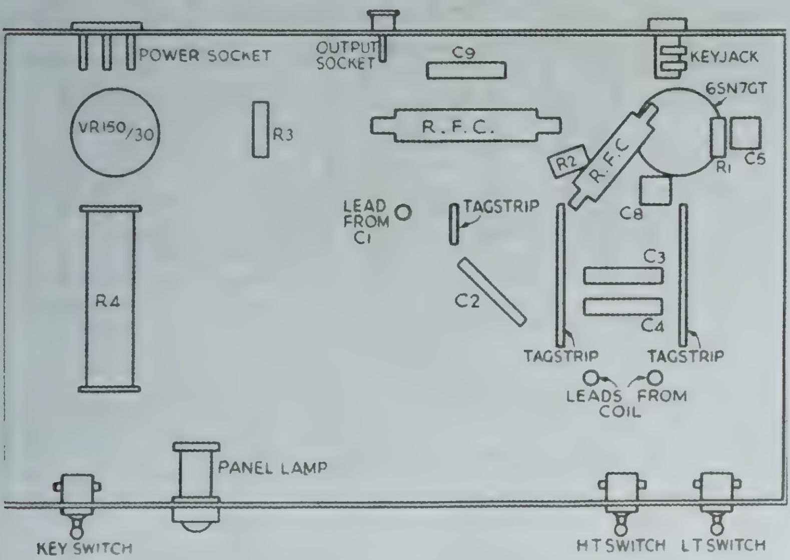


Fig. 16.

Sketch showing layout of components underneath the chassis.

CHAPTER 6 A CRYSTAL-CONTROLLED FREQUENCY SUB-STANDARD

In accordance with the terms of the G.P.O. transmitting licence every amateur station in the United Kingdom must be equipped with means of checking accurately the sending frequency if the actual transmitter is not crystal controlled. The apparatus to be described in this chapter provides a

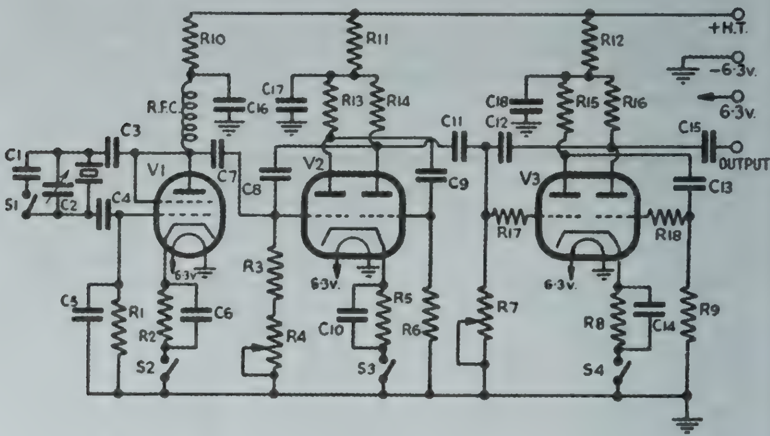


Fig. 17.
Circuit of Frequency Standard giving 1 Mc/s., 100 kc/s. and 10 kc/s. points.

C1	5 $\mu\mu$F. *T.C.C.*, type CC30a.
C2	60 $\mu\mu$F. variable. *Eddystone*, type 582.
C3, 4, 12, 13	·001 μF. *T.C.C.*, type 543.
C5	100 $\mu\mu$F. *T.C.C.*, type CC42h.
C6, 10, 14, 16, 17, 18	1 μF. *T.C.C.*, type CE30N.
C7	2 $\mu\mu$F. *T.C.C.*, type CC30a.
C8, 9	100 $\mu\mu$F. *T.C.C.*, type CC42h.
C11	3 $\mu\mu$F. *T.C.C.*, type CC30a.
C15	15 $\mu\mu$F. *T.C.C.*, type CC30a.
C19, 20	8 μF. *T.C.C.*, type CE5L.
R1	150,000 ohms. *Erie*, type 9.
R2	1,000 ohms. *Erie*, type 8.
R3	30,000 ohms. *Erie*, type 9.
R4, 7	50,000 ohms variable. *Dubilier*, type YN.
R5, 8	150 ohms. *Erie*, type 8.
R6, 9	50,000 ohms. *Erie*, type 9.
R10	25,000 ohms. *Erie*, type 8.
R11	20,000 ohms. *Erie*, type 8.
R12	15,000 ohms. *Erie*, type 8.
R13, 14	10,000 ohms. *Erie*, type 8.
R15, 16	15,000 ohms. *Erie*, type 8.
R17, 18	50 ohms. *Erie*, type 9.
R19	47,000 ohms. *Erie*, type 1.
R.F.C.	*Eddystone* R.F. choke, type 1010.
S1	*Bulgin* press switch, type S422.
S2, 3, 4, 5	*Bulgin* on-off switches, type S261.
V1	*Mullard* EF91.
V2, 3	*Mullard* ECC91.
V4	*Brimar* 6X4.

Other components:
- Chassis. *Eddystone*, type 643.
- Cabinet. *Eddystone*, type 644.
- 3-pin socket. *Belling Lee*, type 1113.
- Valveholders B7G with shields.
- 2-pin mains plug and socket. *Bulgin*, type P74.
- 3 Knobs. *Eddystone*, type 593.
- 3 Condenser spindle extensions. *Eddystone*, type 1008.
- 1 Condenser mounting bracket. *Eddystone*, type 1007.
- 1,000 kc/s. crystal and type M socket. *Q.C.C.*
- 2 Insulated terminals. *Belling Lee*, type L1001/SW.
- T1 *Woden* transformer, type DTM11a.
- L.F.C. *Woden* choke, type DCS11.

simple but extremely effective means of performing this service on any amateur band from 1·7 Mc/s. to 30 Mc/s.

There are, broadly speaking, two methods available to the amateur for frequency checking: a calibrated variable frequency oscillator covering the required range, and a crystal sub-standard provided with means for producing closely-spaced marker points at regular intervals throughout the desired range.

The construction of a variable frequency oscillator to cover the wide bands of frequencies in which an amateur may be interested—and to the required degree of accuracy—is a task not to be lightly undertaken, and together with means for checking its calibration at several points on each range would constitute a formidable undertaking even for an experienced person possessing adequate measuring apparatus.

The second method is that employed in the apparatus to be described. Referring to the circuit diagram in Fig. 17, V1 is a 1 Mc/s. crystal oscillator employing a *Mullard* EF91 pentode connected as a triode, and with the crystal between grid and anode, thus doing away with the necessity for a tuned circuit at the crystal frequency. A suitable crystal may be obtained from the *Quartz Crystal Company* of New Malden, but when ordering it is essential to mention that it is to be used in conjunction with a grid-to-plate Pierce oscillator circuit.

The small variable condenser C2, which is connected in parallel with the crystal, enables the frequency of the latter to be brought exactly into resonance with a standard frequency transmission as described later.

The harmonics of the crystal oscillator are receivable on any normal receiver up to more than 40 Mc/s.

A *Mullard* type ECC91 double-triode valve (V2), is connected as a 100 kc/s. multi-vibrator, and its 10th harmonic, *i.e.* 1,000 kc/s., is locked by the 1 Mc/s. crystal oscillator. The output of a multi-vibrator is extremely rich in harmonics, and marker points are produced at intervals of 100 kc/s., *and of an accuracy equal to that of the 1 Mc/s. crystal* throughout the range 100 kc/s. to at least 40 Mc/s. To assist in the location of the 1 Mc/s. harmonics V2 may be made inoperative when desired by means of switch S3 in its cathode circuit.

With the aid of V1 and V2 the high and low frequency edges of each amateur band can be located, together with several points of reference in each band, but frequencies lying between such points would have to be calculated by interpolation on the receiver tuning scale. To assist in this interpolation the valve V3 is provided. This is also a *Mullard* ECC91 connected as a multi-vibrator and designed to divide the frequency of V2 by 10, so producing reference points separated by only 10 kc/s. These points are sufficiently close together to enable an unknown frequency lying between any two of them to

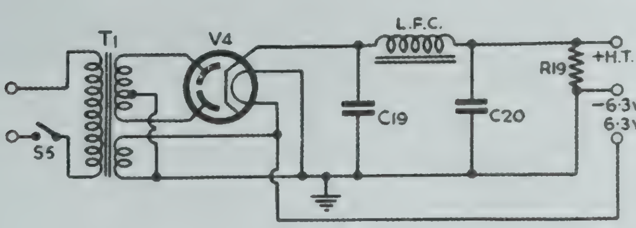


Fig. 18.
Circuit of Pack for frequency standard. See Fig. 17 for details of components.

be estimated to a degree of accuracy which is high enough for all normal purposes. This stage is also provided with a switch, S4, for bringing it into use as required.

Construction

The unit shown in the photograph (page 48) was built on an *Eddystone*, type 643, die-cast chassis and housed in the corresponding cabinet, *Eddystone*, type 644, together with its mains power supply. In order to accommodate all the parts in such a comparatively small space it was found necessary to employ valves and components of small dimensions, but a larger chassis may be substituted. Alternatively the power supply can be housed separately; in such an event larger valves, etc., could be employed. Suitable valves in the octal based range include the *Mullard* ECC32 or ECC33, the *Osram* B65 and the *Brimar* 6SN7GT for the positions V2 and V3, while almost any medium impedance triode, such as the *Osram* L63, or the *Brimar* 6J5 or 6C5, or an R.F. pentode strapped as a triode, would be satisfactory as V1. It must be remembered, however, that although all the valves mentioned are suitable for the duties indicated, they are not necessarily interchangeable without rewiring the valve holders.

While the exact layout of components on the chassis is not of vital importance, grid and anode leads should be kept reasonably short, and in view of the very small values of the coupling condensers C7 and C11, care should be taken that stray capacities, due to an unsuitable disposition of components, do not materially alter their effective values.

While not unduly critical, the values of the resistances and condensers associated with the two multi-vibrators should be reasonably in conformity with those shown in the circuit.

The potentiometers R4 and R7 are operated only during the initial setting up of the circuit and are, therefore, mounted on the chassis behind the front panel, together with C2 (an *Eddystone* type 582 variable condenser) which is mounted on an *Eddystone* bracket bolted to the side of the chassis. Extension rods fitted with small knobs bring these three controls to a convenient position for operation when the lid of the cabinet is raised. Suitable components will be found in the *Eddystone* range (type Nos. 1008 and 593 respectively).

For those who desire to duplicate the original model a sketch of the positions occupied by the major components under the chassis is given in Fig. 20 (page 49).

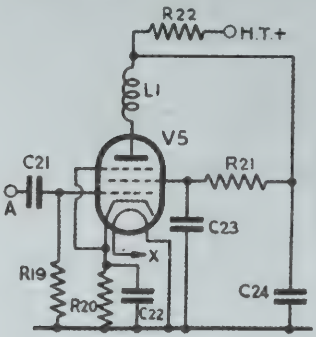


Fig. 19.
Additional stage extending the frequency range to include the 144–146 Mc/s. band. The point marked "A" is connected to the anode of V1.

C21	100 $\mu\mu$F.
C22, 23, 24	500 $\mu\mu$F.
R19	250,000 ohms ½ watt.
R20	22,000 ohms ½ watt.
R21, 22	3,000 ohms ½ watt.
L1	2 Turns 20 S.W.G., ⅜ in. diameter, ½ in. long.
V5	See text.

The Power Supply

As the H.T. and L.T. requirements of the frequency standard are quite modest, the designer of the prototype decided to make the power supply available for energising other apparatus in use at the station irrespective of whether the frequency standard is in operation or not. To this end the H.T. positive and negative and the " live " L.T. connections are brought out to a *Belling Lee* three-pin socket at the rear of the chassis. The *Woden* mains transformer and choke occupy positions to the left and right at the rear of the chassis when viewed from the front, the *Brimar* 6X4 miniature rectifier valve being between them. Reservoir and smoothing condensers are provided by an 8 – 8 μF. unit of *T.C.C.* manufacture. It will be noted that the 6X4 rectifier derives its heater voltage from the common 6·3 V. winding: this is possible as this valve, like its larger counterpart the 6X5, possesses adequate heater cathode insulation to withstand the H.T. voltage.

Setting up the Frequency Standard

With S4 in the " off " position, the crystal oscillator stage only should be switched on by operating the switch S2–S3. With the beat frequency oscillator of a receiver in operation two adjacent harmonics should be located and their dial readings noted. To minimise pick up of unwanted signals and in order that the harmonics may be readily identified, the aerial should be removed from the receiver and a short length of wire from the aerial terminal run near or actually connected to the *Belling Lee* terminal on the unit marked " output."

The 100 kc/s. multi-vibrator should now be brought into use by means of the switch S3, and a careful search made between the two harmonic points just obtained. A number of " pips " should be heard, and if the potentiometer R4 is rotated these will change their frequency in a jerky manner, interspersed with an untunable rushing noise. Count the number of " pips " *between* the two crystal harmonics, and if there are more or less than *nine* adjust R4 until this condition is satisfied. Next tune the receiver to one of the " pips," or 100 kc/s. marker points, and move R4 slowly first one way and then the other until the multi-vibrator jumps out of lock. R4 should be left set in a position mid-way between the positions where this occurs.

To align the 10 kc/s. multi-vibrator, tune-in two of the adjacent 100 kc/s. points, and make a note of the dial reading of each. This should be done on a frequency range on the receiver where the band-spread is sufficient for the two points to be well separated. The second, or 10 kc/s., multi-vibrator should now be brought into circuit by operating switch S4. A further series of marker " pips " should again be heard between the two 100 kc/s. positions. Potentiometer R7 should then be adjusted in exactly the same way as was R4 until *nine* pips are heard *between* any two adjacent 100 kc/s. points. The setting of R7 is rather more critical than that of R4, but not unduly so, and after it has been adjusted to the mid position as previously described the instrument is ready for its final checking.

Comparison with a Standard Frequency Transmission

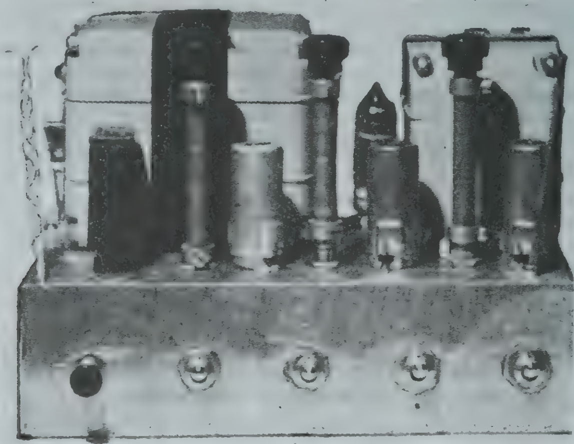
This operation may be conveniently carried out by comparing the frequency of a harmonic of the crystal with one of the transmissions from WWV, the station of the American Bureau of Standards in Washington. This station is well received in the United Kingdom at almost any time of the day or night. A list of the frequencies radiated by WWV is given on page 30.

WWV should be tuned in on the receiver with the beat frequency oscillator switched off, and the 1 Mc/s. crystal oscillator in the frequency standard switched on. The strength of the local signal should be comparatively weak so as not to overload the receiver. When condenser C2 is moved slowly, a position will be found where the harmonic of the crystal is exactly zero-beat with WWV. If the maximum accuracy of which the frequency standard is capable is required, considerable care should be exercised in arriving at this adjustment. It should, however, be understood that with C2 set at minimum capacity the frequency of the crystal will be accurate enough for all normal measurements.

Operation

Assume that a frequency, known to lie between 7 and 8 Mc/s., is to be measured. First locate the 7 Mc/s. harmonic of the crystal with V1 alone operative. Now bring in the 100 kc/s. multi-vibrator by means of switch S3, and count the number of 100 kc/s. points between 7 Mc/s. and the unknown frequency. Assume there are two. This places the unknown somewhere between 7·2 and 7·3 Mc/s. Tune the receiver to 7·2 Mc/s. and bring into circuit the 10 kc/s. multi-vibrator by operating switch S4, counting the number of 10 kc/s. points which lie between 7·2 Mc/s. and the unknown frequency. Should there be three, the required frequency is now known to be between 7,230 and 7,240 kc/s., and interpolation on the dial of the receiver will enable an estimate, to at least the nearest kilocycle, to be made provided the receiver band-spread is adequate.

The *Bulgin* push-button switch, S1, is provided to enable signals from the frequency standard to be easily identified at times when other signals are present in the receiver. The effect of bringing S1 into circuit is to move



Front view of the chassis. The signal-identification switch (S1) is on the left of the front apron, with switches S2 to S5 to the right. The flex on the left and the insulated lead on the right are connected to the dial lamp and output terminal respectively, both components being mounted on the front of the cabinet.

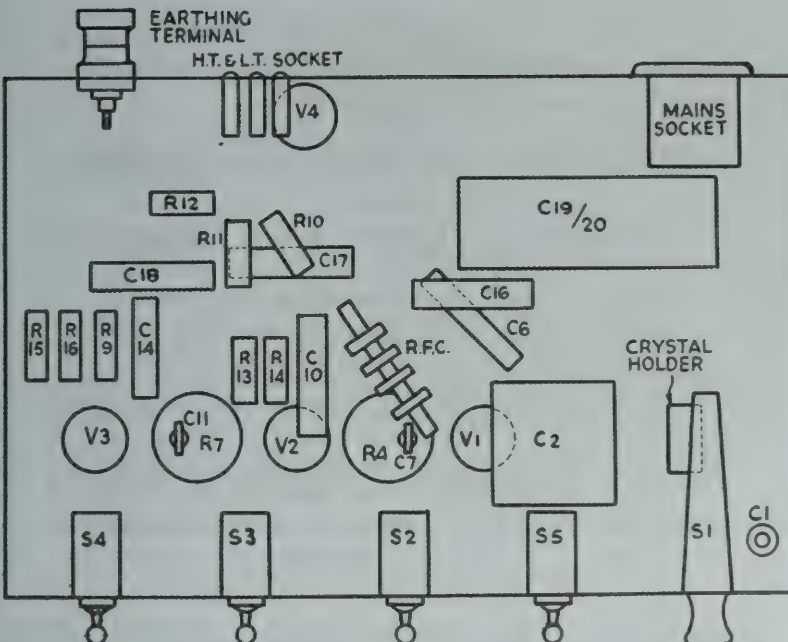


Fig. 20.
Sketch plan of the position of the major components under the chassis.

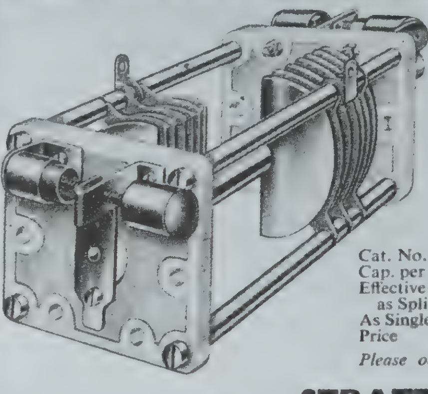
all the marker points downwards in frequency by a small amount. This is accomplished by connecting C1 temporarily in parallel with the crystal. S1 should *not* be held down when making an actual measurement.

The Frequency Standard on V.H.F.

Using a sensitive receiver 10 kc/s. marker points up to more than 60 Mc/s. have been located. To extend the range to 144 Mc/s. an additional valve stage may be added, the circuit for which is given in Fig. 19. On the 144 Mc/s. band, both the 1 Mc/s. and the 100 kc/s. harmonics are available at ample strength.

None of the component values is at all critical, and output may be taken either from a single-turn link closely-coupled to L1, or through a 5 $\mu\mu$F. condenser connected from the anode of V5 direct to the input of the receiver.

If it is desired to employ a miniature valve, V4 may be a *Mullard* EF91, or if space permits an EF50 or an EF54, by the same makers, will be equally suitable. There are no controls associated with this stage. The H.T. requirements are extremely small, and as it has no detrimental effect upon performance on the lower frequencies, there is little point in making provision for switching it in or out of circuit. The grid drive is taken from the anode of V1, and a convenient point at which to pick-up the H.T. line is at the junction of R10 and the R.F. choke in Fig. 17.



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